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Range extension in acoustic distance measurement.

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Range Extension in Acoustic Distance Measurement

by

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Abstract

This work presents the application of chirp techniques used commonly in microwave radars to the problem of ultrasonic ranging in the atmosphere. The object of this thesis was to extend the range of the commercially available Polaroid ranging system, which according to the manufacturer's specifications has a maximum range of 35 feet or 11.7 meters.

Linear frequency modulation of a sinusoid at the transmitting end results in a signal with a large bandwidth content, which combined with matched filtering at the receiver produces the effect of pulse compression on the received signal. This is specifically important in ranging systems as the operator is only interested in ascertaining as to whether a replica of the transmitted signal is present in the signal is present in the echo or not. Amplitude weighting of the matched filter impulse response to reduce the range sidelobes in the matched filter output has been investigated.

The circuitry provided by Polaroid has been modified in order to enable ranging to larger distances. The processing station was an IBM compatible 486 machine, with a signal processing board on it. The software used for the purpose was 'SPOX' which is C language type digital signal processing development system.

To Uma chitthi and Raju chitthappa for everything.

ACKNOWLEDGMENTS

I would like to thank my supervisor Dr. J.J.Soltis for his support and guidance throughout the course of this work.

Thanks to all my friends for their help and encouragement, especially Rama and Majety.

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NOMENCLATURE

θ	Azimuthal angle about the transducer
f_0	Center frequency of the Chirp signal
k	Chirp rate
$\varepsilon_1(t)$	Complex Chirp signal
$R_{ab}(\tau)$	Cross Correlation function between a and b
D	Dispersion Factor
T	Duration of signal
ξ	Energy of Signal
f	Frequency
Δ	Frequency Sweep (or Bandwidth) of the Chirp Pulse
$h(t)$	Impulse response of filter
f_i	Instantaneous frequency
N	Length of Filter
L_s	Loss Factor
A_{max}	Maximum attenuation
$E(x)$	Mean Value of x
$n(t)$	Noise component at input of Matched Filter

M	Number of data Samples
B_n	Noise Equivalent Bandwidth
$y_o(t)$	Output of Matched filter
φ	Phase
$\eta/2$	Power spectral Density of White Noise
R	Range to Target
P_r	Received Power
$s(t)$	Signal component at input of Matched Filter.
t	time
a	Transducer Radius
P_{Tr}	Transmitted power
c	Velocity of propagation of ultrasound
λ	Wave length
rect(z)	Rectangular function
Rx	Receiver
SNR	Signal to Noise Ratio
TEMP	Ambient Temperature in degrees Kelvin
Tx	Transmitter

Var(X)	Variance of X

I. Introduction

There are many different ways of acquiring range information. These techniques can easily be divided into two groups, passive and active range detectors. Active range detectors operate by transmitting a signal, acquiring the range information by measuring the return signal. Passive detectors provide range information through computationally intensive procedures like geometric manipulations.

Active range detectors are classified by the type of signal they emit , either electromagnetic (e.g. radar) or ultrasonic (e.g. sonar). For measuring short distances the velocity of an electromagnetic signal makes the measurement of time of flight very difficult. Laser ranging devices are commercially available but they tend to be expensive. While electromagnetic ranging is excellent for long range applications, ultrasonics is more suitable for short ranges. Because sound absorption in water can be very low compared to acoustic transmission in air and electromagnetic transmission in water, sonars are generally used in underwater applications.. There exists another class of systems that makes use of both ultrasonic and electromagnetic waves to perform the required ranging[Daas, Knochel].

Ultrasonics are finding increasing use in robotics research as a means of ranging and obstacle avoidance and map building [Michael Brown]. They have also found varied other uses like acoustic pyrometry, level measurement, biomedical applications, nondestructive testing and atmospheric echo sounders.

Due to the rapid attenuation in air with distance, ultrasonic ranging is not very popular at distances greater than a few meters. The objective of this investigation is to study the possibility of extending the range of the commercially available ultrasonic transducer manufactured by Polaroid by using techniques similar to those used in radar systems.

II. Theoretical Background

2.1 Ultrasonic Transducers

Ultrasonic ranging (Sonar) and detecting devices use high frequency sound waves to detect the presence of an object and its range. The systems either measure the echo reflection of the sound from objects or detect the interruption of the sound beam as the objects pass between the transmitter and receiver. A transmitting transducer sends out a pulse of sound that is detected by a receiving transducer[Shirley]. Figure 2.1 shows several different transducer configurations. In figure 2.1(A) the transmitter and receiver are placed side by side. The receiver detects the energy reflected by the object. Figure 2.1(B) shows the setup when the same transducer acts as both the transmitter and receiver. The above describes the transducer setup used primarily in ranging applications. Figure 2.1(C) shows the setup that could be used for applications like counting and alignment. Here the transmitter and receiver are mounted opposite each other.

Ultrasonic transducers are often designed to be directional so that the sound is efficiently transmitted or received only over a certain conical beam in front of the transducer. Transducers can be designed that vary from being highly directional to omni-directional. A transducer with a narrow beam pattern will only detect those objects that lie in the insonified path of the transducer. Also narrow beam angle systems are less susceptible to background ultrasonic noise and also have a greater range.

The Polaroid ultrasonic transducer is an electrostatic type. It is composed of a very thin Kapton film diaphragm, vacuum coated with gold to form the negative electrode. The positive electrode is an aluminum back plate which also provides the resonant structure for the diaphragm. The transmit and receive responses are quite flat in the range 20-100 kHz. The Transducer is about 1.5 inches in diameter and has 3 dB full angle beam width of about 15 degrees.

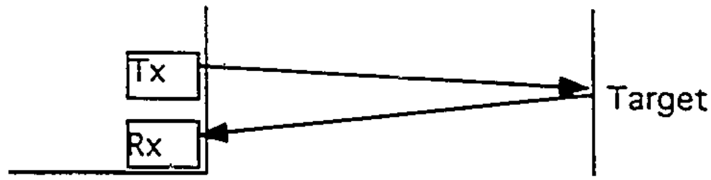

Figure 2.1(A) Echo of the Ultrasonic Beam is reflected from target to a Receiving Transducer

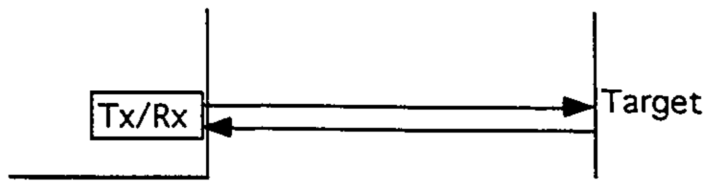

Figure 2.1(B) One Transducer both Transmits and Receives the Beam.

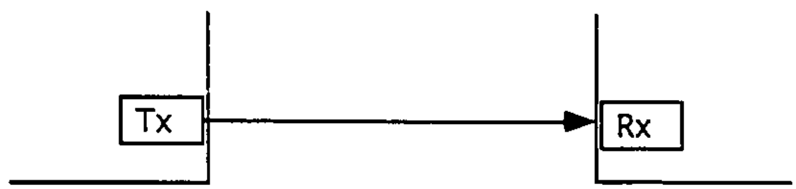

Fig 2.1(C) Transducers mounted opposite each other

The major acoustical factors affecting the performance of an ultrasonic ranging system are related to the transducer performance, operating frequency and the desired maximum range. There are two major relationships that affect the performance. The first is the relationship between transducer size, beam width and operating frequency.

The transducer can be treated as a plane circular piston set in an infinite baffle. Its radiation pattern $P(\theta)$, which is the radiated power as a function of direction is then given by :

$$P(\theta)=\frac{2\mathrm{J}_1(ka\sin(\theta))}{ka\sin(\theta)} \qquad (2.1)$$

where k = wave number = $\frac{2\pi}{\lambda}=\frac{2\pi f}{c}$, a = piston radius , θ = azimuthal angle. $\mathrm{J}_1(x)$ is a Bessel function of the first kind, of order one, and argument x. The beam width is mostly expressed in terms of the angle intercepted by the points either side of the principal axis where the radiation pattern is 3 dB less than the on axis value (θ=0). Thus the right hand side of 2.1 can be set as

$$\frac{2\mathrm{J}_1(ka\sin(\theta))}{ka\sin(\theta)}=\text{-3dB}=\frac{1}{\sqrt{2}} \qquad (2.2)$$

Expanding the Bessel function to 3 terms yields

$$ka\sin(\theta)=1.62 \qquad (2.3)$$

The full 3 dB beam width is thus given by

$$2\theta=\sin^{-1}(1.62/ka) \qquad (2.4)$$

The second relationship is between range and operating frequency. The maximum range of a ranging system is a function of its frequency due to the frequency dependence of the attenuation of sound in air. The maximum attenuation for ultrasonic sound ignoring the effects of temperature and humidity can be approximated as

$$A_{max}=3.3*f*10^{-2}\ \text{dB/meter} \qquad (2.5)$$

where f = the frequency in kHz.

Thus attenuation limits the range of higher frequency transducers. However background noise at the higher frequencies is less and higher frequency ultrasonic transducers have a better chance of operating in acoustically noisy environments than do lower frequency transducers[Shirley].

The velocity of sound in air is greatly affected by the temperature. The velocity of sound in air at any temperature is given as

$$c = 331.4 * \sqrt{\frac{\text{TEMP}}{273}} \text{ meters/second} \qquad (2.6)$$

where TEMP is the ambient temperature in degrees Kelvin.

2.2 Time Delay Estimation For Active Ranging

If a transmitter generates an electrical signal $S_T(t)$, which excites an ultrasonic transducer, the resulting acoustic waves radiate outward towards a reflecting target. When the waves strike the target, they are reflected and returned to the transducer. The transducer then converts the reflected wave to an electrical signal $S_R(t)$. If the interfering ambient noise and the noise generated by the receiver are assumed to be additive then the electrical signal at the receiver is given by

$$y(t) = S_R(t) + n(t) \qquad (2.7)$$

where $n(t)$ is electrical signal that represents the combined additive noise. In general $S_R(t)$ depends on the physical properties and the shape of the reflecting target, $S_T(t)$ and the response of the transducer.

The time delay estimation problem is to determine the time delay between the time $S_T(t)$ was transmitted, say t=0 and the time $S_R(t)$ was received say $t=\tau$. In the absence of additive noise the estimation of the time delay is not difficult as detecting the leading edge of $S_R(t)$ is quite easy. However in the presence of additive noise detecting the leading edge of $S_R(t)$ becomes more difficult. This difficulty increases as the signal to noise ratio decreases. This means that success in determining the time delay depends on careful design of the transmitted signal and on the received Signal to Noise ratio[Elliot]. Figure 2.2 shows the block diagram for the time delay estimation problem.

In general the signal to be extracted from an interference background often consists of just a few valid data points as compared to amount of data to be processed. The case of the time delay estimation problem too falls under this class.

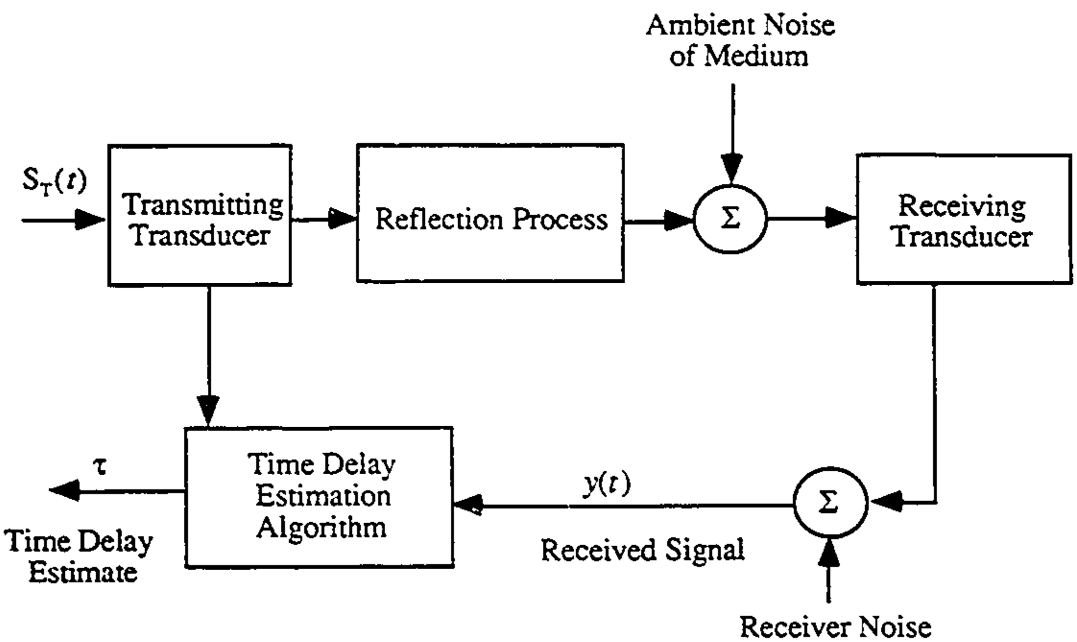

Figure 2.2. Block Diagram of the Time Delay Estimation Problem.

In order to perform the ranging operation it is first necessary to illuminate the target with sufficient energy so that there is some minimum reflected signal energy to process. In the case of pulsed Radars and Sonars the range is limited by the average power transmitted and in resolution by the pulse width. The design of simple pulsed systems involves a compromise between these two factors.

Good range resolution demands a very narrow pulse, long range requires that the amount of power transmitted be large, which in turn implies transmitting a long pulse due to the limitations of peak instantaneous transmitter power. It is the large bandwidth that a short pulse possesses that gives it the property of good range resolution. A long pulse at a constant carrier frequency contains a narrow bandwidth and hence poor range resolution. The spectrum of a long pulse can be widened by introducing modulation. Using frequency modulation of the carrier, the frequency spread characteristics of a short pulse can be introduced in a long duration signal.

III. Fundamentals of Chirp and Matched Filters

3.1 Chirp

As discussed earlier, a desirable signal for a pulsed ranging system would possess a large bandwidth and transmitted power. Frequency modulating the carrier is one method of achieving the above. The class of signals defined by the real part of equation 3.1(a) are called as CHIRP signals[Darlington]. Equation 3.1(b) gives the real part of 3.1(a).

$$\varepsilon_1(t) = \mathrm{rect}(\frac{t}{T})e^{2\pi j(f_0 t + kt^2)} \tag{3.1(a)}$$

$$\mathrm{S}(t) = \mathrm{rect}(\frac{t}{T})\cos(2\pi(f_0 t + \frac{kt^2}{2})) \tag{3.1(b)}$$

where f_0 indicates some suitable carrier frequency, T is the signal duration, k is a constant called the CHIRP rate and

$$\mathrm{rect}(\mathrm{z}) = \begin{matrix} 1 \text{ if } |\mathrm{z}| < 0.5 \\ 0 \text{ if } |\mathrm{z}| > 0.5 \end{matrix} \tag{3.2}$$

The signal defined by equation 3.1 is taken to be of unit amplitude.

The phase of the signal is given by

$$\varphi = 2\pi(f_0 t + \frac{kt^2}{2}) \tag{3.3}$$

The instantaneous frequency is defined as

$$f_i = \frac{1}{2\pi}\frac{d\varphi}{dt} = f_0 + kt \tag{3.4}$$

Thus during the T second interval of the pulse the instantaneous frequency changes in a linear fashion from $f_0 - \frac{kT}{2}$ to $f_0 + \frac{kT}{2}$. The net frequency sweep, Δ is then given by the difference of the two values or kT. This implies that the amount of frequency sweep is symmetrical about the time $t=0$. This objection can be avoided by adding a sufficiently large constant time delay such that the time is positive for all frequencies of interest. A dimensionless product frequently appears in the case of chirp signals and is called the DISPERSION FACTOR which is given by:

$$D = T\Delta \tag{3.5}$$

The signal as described by equation 3.1 possesses the characteristics as shown in figures 3.1(A) and 3.1(B).

The larger the value of the dispersion factor D, the spectrum of the Chirp signal becomes more nearly rectangular, with a total bandwidth approaching Δ. For smaller values of D there is considerable frequency content outside of a band Δ centered at f_0. However almost 95% of the spectral energy is contained in the band Δ even for Dispersion factors as low as 10.

The transmitted signal would be reflected by the target back towards the transmitter. The received echo contains the range data in the sense that the time taken for the signal to travel to the target and back is directly proportional to the distance (this may not always be the case, especially in environments where multipath reflections occur) between the transmitter and the target. Thus the time delay has to be extracted from the received signal. One way of doing this is have a receiving network with delay vs. frequency characteristics as shown in figure 3.2. This is basically just the time reversed version of the fig 3.1(B).

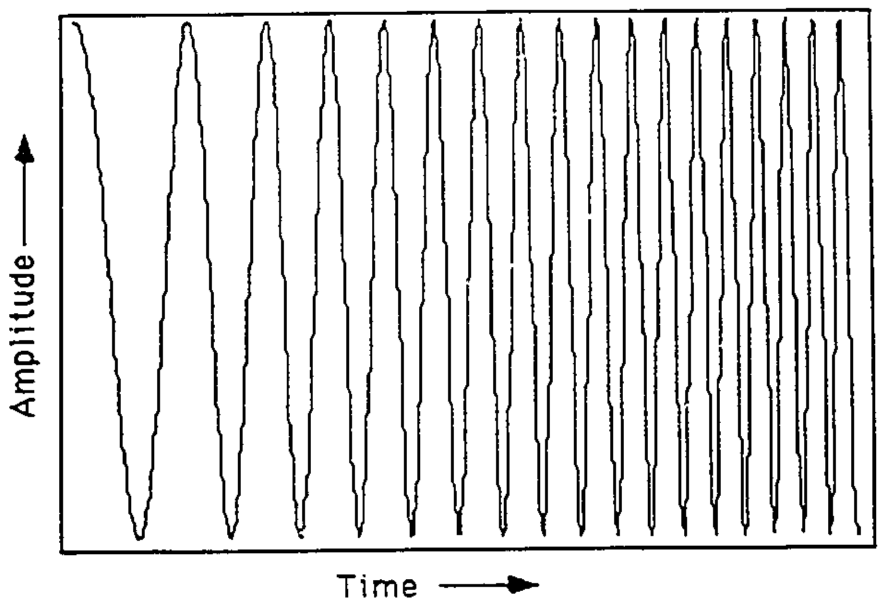

Figure 3.1(A) Schematic Diagram of a Chirp

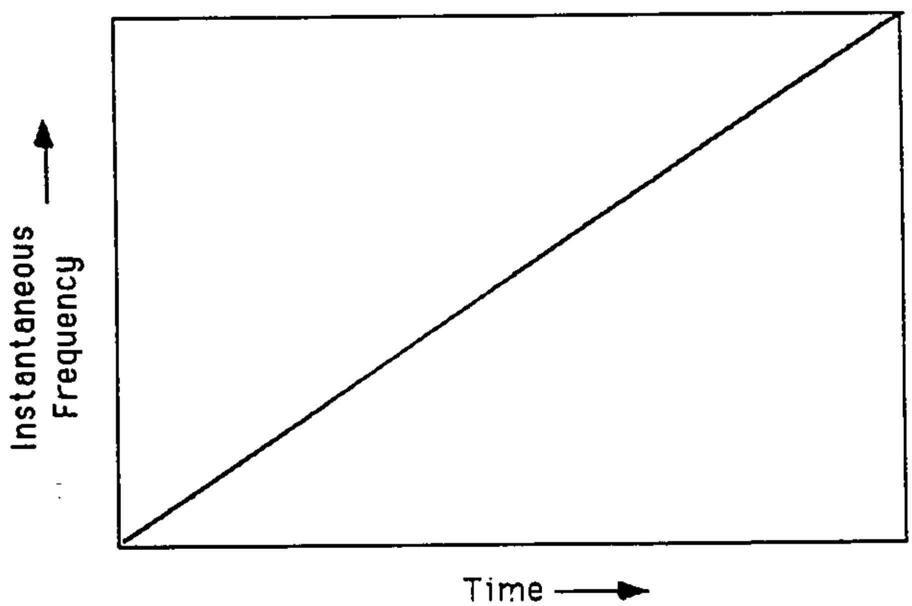

Figure 3.1(B). Instantaneous Frequency vs. Time Characteristic of a Chirp Signal.

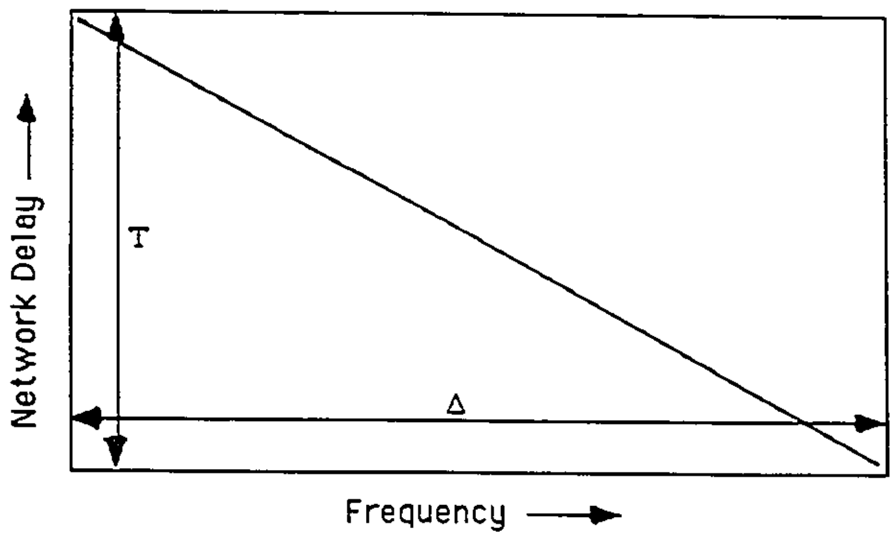

Figure 3.2: Delay vs. Frequency characteristics of the Receiving network.

3.2 Matched Filter

Measuring the round trip delay of the transmitted signal to the target and back is the basis of all pulsed ranging systems. In most cases, however, the amount of reflected power is very small. The Matched Filter is the filter that maximizes the received Signal to Noise ratio and hence is an optimum processor in the sense that it maximizes the probability of detecting a signal buried in noise. This is very important in ranging applications as one is essentially dealing with detection of the echo reflected by the target.

If $s(t)$ is any physical waveform, then a filter that is matched to $s(t)$ is defined as one with an impulse response given by :

$$h(\tau) = \begin{cases} k_s s(T-\tau) & 0 \le \tau \le T \\ 0 & elsewhere \end{cases} \qquad (3.6)$$

where k_s is an arbitrary scale factor and can be taken to be unity[Turin]. The transfer function of a matched filter, which is the Fourier transform of the impulse response is the complex conjugate of the Fourier transform of the signal to which it is matched. For this reason it is also called a CONJUGATE filter. The impulse response is the mirror image of the signal to which the matched filter is matched. It should be noted here that matched filters are used for signal recognition in the presence of noise, and not for signal fidelity, S which requires a flat frequency response.

The output of a matched filter in the presence of additive noise $n(t)$ is given by:

$$y_o(t) = \int_0^T s(T-\tau)s(t-\tau)d\tau + \int_0^T s(T-\tau)n(t-\tau)d\tau \qquad (3.7)$$

where the signal component of $y_o(t)$ is given by the first term in equation 3.7 and reaches its maximum at $t = T$. The output Signal to Noise Ratio (SNR) is given by:

$$\mathrm{SNR}_0 = \frac{\{E_S[y_o(T)] - E_{NS}[y_o(T)]\}^2}{\mathrm{var}_S[y_o(T)]} \tag{3.8}$$

where $E[.]$ stands for the mean value and var[.] stands for the variance and the subscripts S and NS indicate if the first term in 3.7 is present or absent. When the interference is white with power spectral density $S_n(\omega) = \eta/2$ the output SNR is given by:

$$\mathrm{SNR}_o = 2\xi/\eta = 2T\Delta(\mathrm{SNR}_i) \tag{3.9}$$

where SNR_i is the input SNR and ξ is the signal energy and is given by

$$\xi = \int_0^T s^2(t)dt \tag{3.10}$$

From equation 3.9 it is obvious that all signals with the same energy perform equally well, irrespective of their time bandwidth products. However the need for large time bandwidth product signals stem from the following considerations:

1) Limitations in the peak signal power which leads to spreading the energy over a larger time period.

2) From the point of view of detection, in the face of white Gaussian noise all signals with the same energy are equally effective, while in the case of band limited interference then of all the signals with the same energy, those with the largest bandwidth are the most effective.

3) Overcome topographical anomalies that might cause a single frequency pulse to be canceled by a reflection.

3.3 Matched Filter for a Chirp

For the Chirp signal in equation 3.1 the corresponding matched filter has an impulse response given by:

$$h_{m1}(t) = \mathrm{Real}(\mathrm{rect}(\frac{t}{T})e^{2\pi j(f_0 t - kt^2)}) \qquad (3.11)$$

The output response when the chirp signal is passed through its own matched filter is given by:

$$y_0(t) = \int_{-\infty}^{+\infty} \mathrm{rect}(\frac{\tau - t}{T})\mathrm{rect}(\frac{\tau}{T})e^{2\pi(f_0 t + (k/2)\tau^2 - (k/2)(\tau - t)^2)}d\tau \qquad (3.12)$$

When $0 \le \tau \le T$ the above becomes

$$y_0(t) = \frac{1}{\pi kt}e^{2\pi j f_0 t}\sin(\pi(ktT - kt^2)) \qquad (3.13)$$

The output envelope of equation 3.13 for $|t| \le T$ is given by:

$$\frac{T}{\pi\Delta|t|}\sin(\pi(\Delta|t| - kt^2)); \qquad (3.14)$$

for $|t| > T$ the envelope vanishes. Thus the output envelope of the compressed pulse produced by passing a chirp signal through its own matched filter very closely approximates the sampling function especially for large values of the Dispersion Factor. The matched filter for the chirp does the operation of pulse compression on the received echo from the target. Thus the output of a matched filter is not a replica of the input signal.

The output of the matched filter is proportional to the input signal cross-correlated with a replica of the transmitted signal, except for the time delay. The cross-correlation function of two signals $f(t)$ and $g(t)$ is given by :

$$R_{fg}(\tau) = \int_{-\infty}^{\infty} f(t)g(t+\tau)dt \tag{3.15}$$

The replica of the transmitted signal is built into the matched filter by means of the impulse response. The filtering operation is itself a convolution of the signal to be filtered and the filter impulse response. Convolution can be thought of as a correlation operation with one of the inputs reversed. In matched filtering the impulse response of the filter is the time reversed version of the signal it is matched to. When the signal is convolved with the impulse response, either the signal or the impulse response has to be reversed in time to perform the filtering. Thus this essentially becomes a correlation operation. If the input signal to the matched filter were the original signal itself then the output of the matched filter would be the auto-correlation function.

The idea of pulse compression is to operate the ranging system with long pulses to obtain the resolution and accuracy of a short pulse but the detection capability of a long pulse. The pulse compression filter speeds up the higher frequencies at the trailing edge of the chirp relative to the lower frequencies at the leading edge. The result is that the energy contained in the original pulse of duration T is compressed into a shorter pulse of duration . If the time delay through the matched filter is too short, compression will not be complete. On the other hand if the delay is too large the compressed pulse width will become broader. The amount of compression that can be achieved depends on the bandwidth of the signal.

In the preceding discussion the output envelope has a minimum side lobe level of about 13.5 dB below the maximum of the main lobe. The remaining sidelobes decay monotonically and symmetrically about the main lobe. In situations with multiple echoes, each of the returning echoes would produce a response envelope similar to that in Figure 3.5(a) but with different amplitude levels. Thus the sidelobes produced by a strong echo could potentially mask the mainlobe of a much weaker echo. In this case the relative levels of the sidelobes would have to be decreased in order to enhance the detectability of weak echoes.

3.4 Generation Of Chirp Signals

Figure 3.3 shows one method of generating a Chirp signal. This is the so called Passive generation method. Here a short pulse of duration approximately equal to that of desired compressed pulse (output of matched filter) is fed to one end of the tapped delay line. The taps are located at uniformly spread intervals along the delay line. A weighting function is applied to each of the taps and the taps are combined into a single output. The width of the combined signal will be equal to the time T that it takes for the pulse to travel the length of the line. The matched filter in this case would be similar to the generator but with the taps in reverse order.

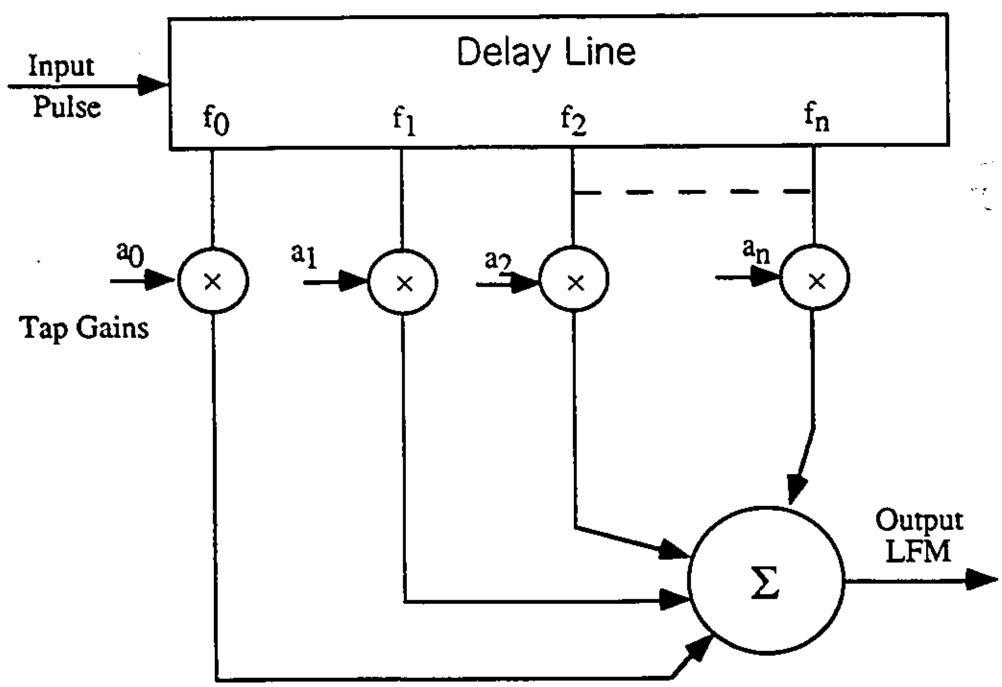

Figure 3.3. Tapped Delay Line Chirp Generator and Matched Filter.

3.5 Sidelobe Reduction in the Matched Filter Output

For a uniformly weighted Chirp the first sidelobe is approximately 13.5 dB below the main lobe peak. By amplitude weighting of the frequencies the side lobe levels can be significantly reduced. However in doing this there is an inevitable broadening of the main lobe and output SNR degradation. The reason for the mainlobe widening is due to the Fourier transform reciprocity relationship. When the spectral shape is made narrower, the time domain waveform becomes broader and vice versa [Oppenheim].

Amplitude weighting can be achieved by the use of window functions that find wide use in the design of Finite Impulse Response (FIR) digital filters. In the case of FIR filters, window functions are applied by arbitrarily truncating an infinite duration impulse response in order to provide an impulse response that is of finite duration. In this case the frequency response of the actual filter would be the convolution of the frequency response of the ideal filter and the Fourier transform of the window. This would result in the actual frequency response being a smeared version of the desired response. For the case of a low pass FIR filter this phenomenon is illustrated in Figures 3.4. Figure 3.4(a) shows the desired frequency response of the lowpass filter. Figure 3.4(b) shows the Fourier transform of the rectangular window function and figure 3.4(c) show the frequency response of the resulting filter.

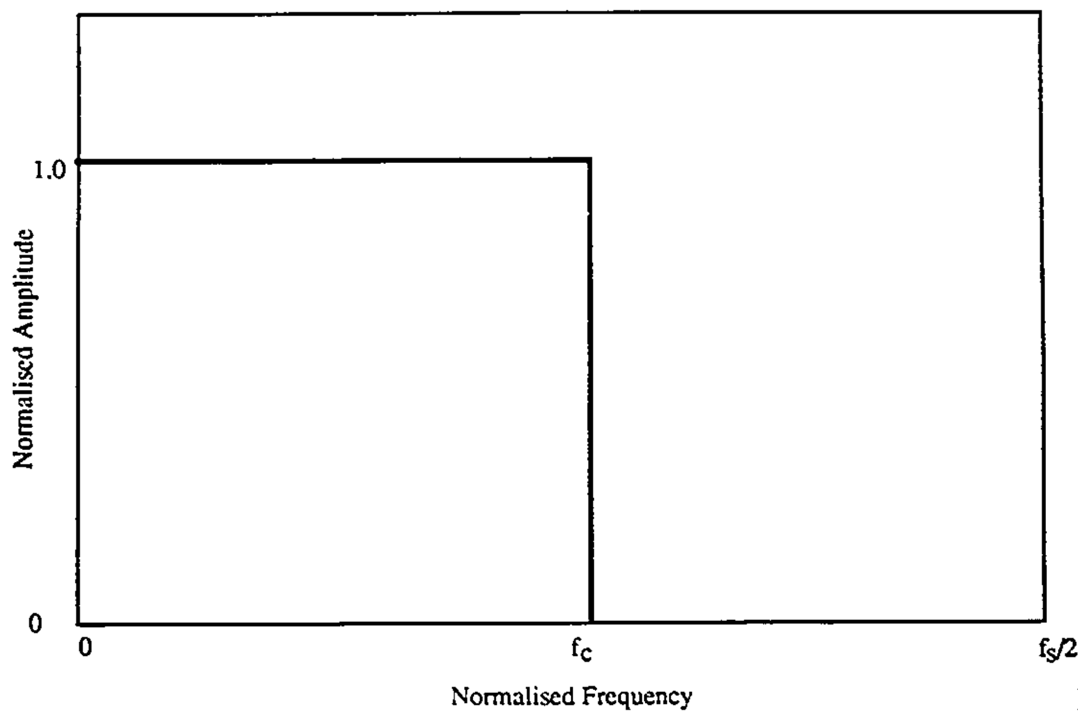

Fig 3.4(a) The desired Frequency Response of a Lowpass Filter.

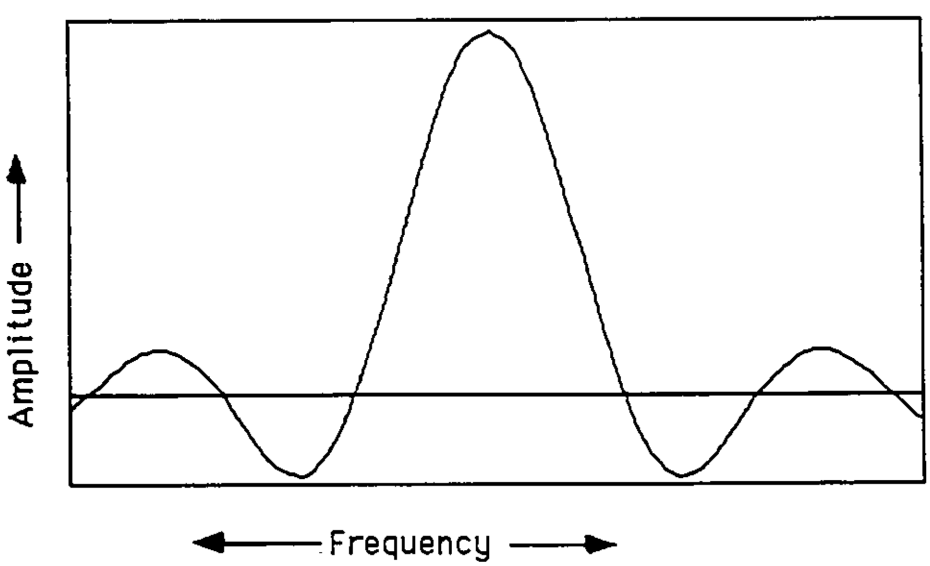

Figure 3.4(b) Fourier Transform of a Rectangular Window.

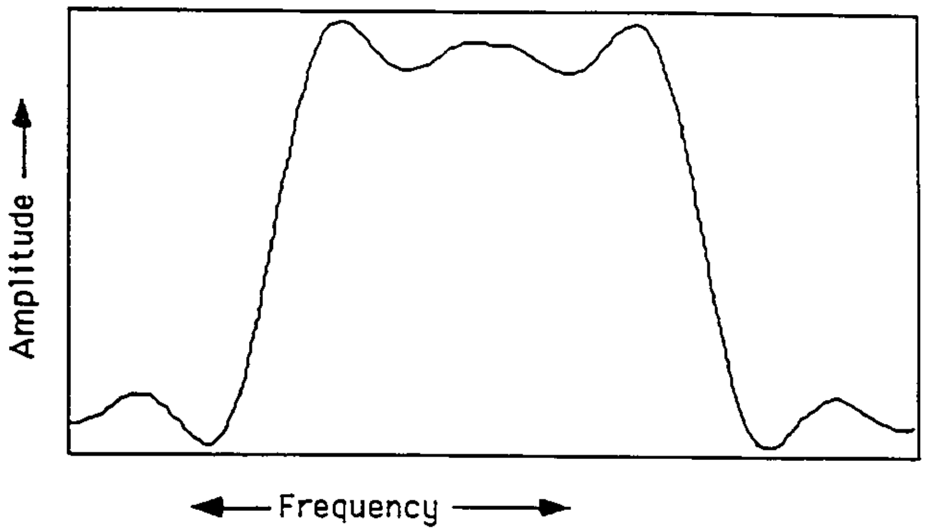

Figure 3.4(c) Frequency response of the resulting filter.

When applying a rectangular window, as the order of the window function is increased the peak amplitudes of the main lobe and sidelobes in its Fourier transform also increase in such a manner that the area under each lobe remains the same while the width decreases. As each lobe in the Fourier transform of the window function slides by a discontinuity in the desired frequency response of the filter during the convolution operation, the result of the convolution will vary in an oscillatory manner. This phenomena is called Gibbs oscillations. Increasing the order of the window does not reduce the amplitude of the oscillations.

In order to reduce the sidelobes of the output of the matched filter for the chirp signal similar window functions have to be applied to the filter coefficients of the matched filter. The idea in this case is not to truncate the length of the filter, rather it is to reduce the levels of the frequency components at the leading and trailing edges of the chirp pulse. Thus the center frequency of the filter has the highest weight while those components at the edges have less weight. The process of applying a window to the Matched filter in effect results in a mismatched filter with the resulting degradation in the signal to noise ratio.

Figures 3.5(a) and 3.5(b) show the output envelope and the log magnitude plot of the same for a matched filter without any amplitude weighting. The chirp pulse had a dispersion factor of 52.2, pulse duration 5.11 milliseconds and a center frequency of 35110 Hz. Although the output has 1024 points only the middle 256 points centered around the main peak are shown. It is obvious from Figure 3.5(b) that the minimum side lobe level is about 13.5 dB's below the mainlobe. Figures 3.6 and 3.7 are for the same signal, but with Hamming and Bartlett windows applied to Matched filter impulse response.

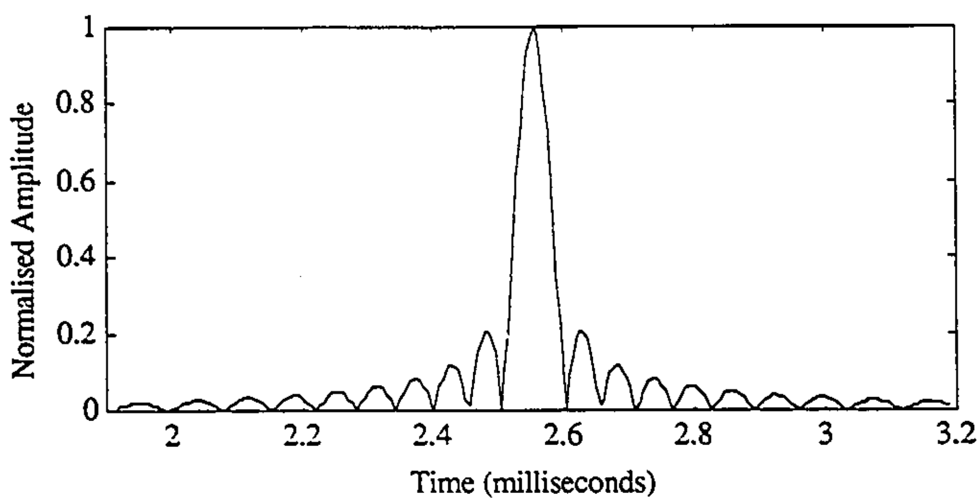

Figure 3.5(a). Output envelope of an unweighted Matched Filter for a Chirp.

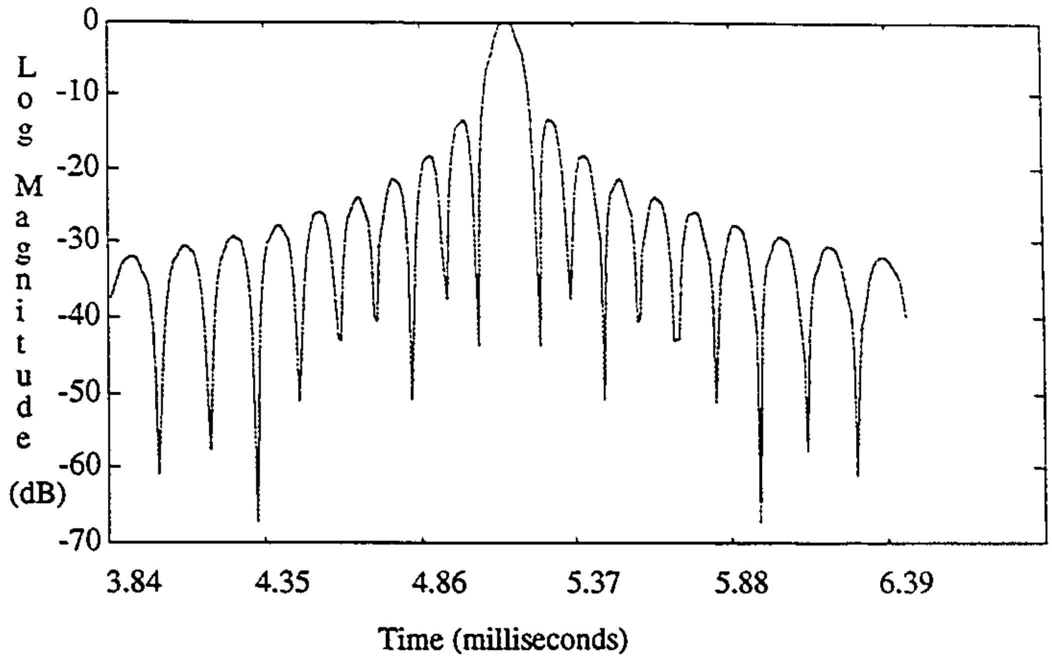

Figure 3.5(b) Log Magnitude plot of the output envelope.

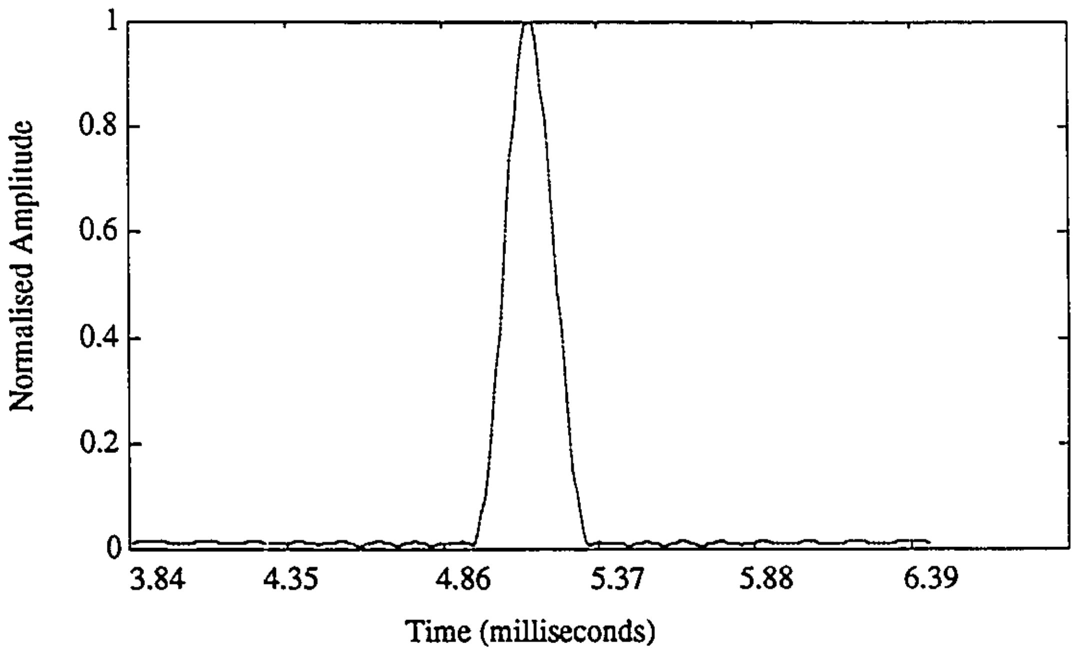

Figure 3.6(a) Output envelope of Filter using Hamming weighting.

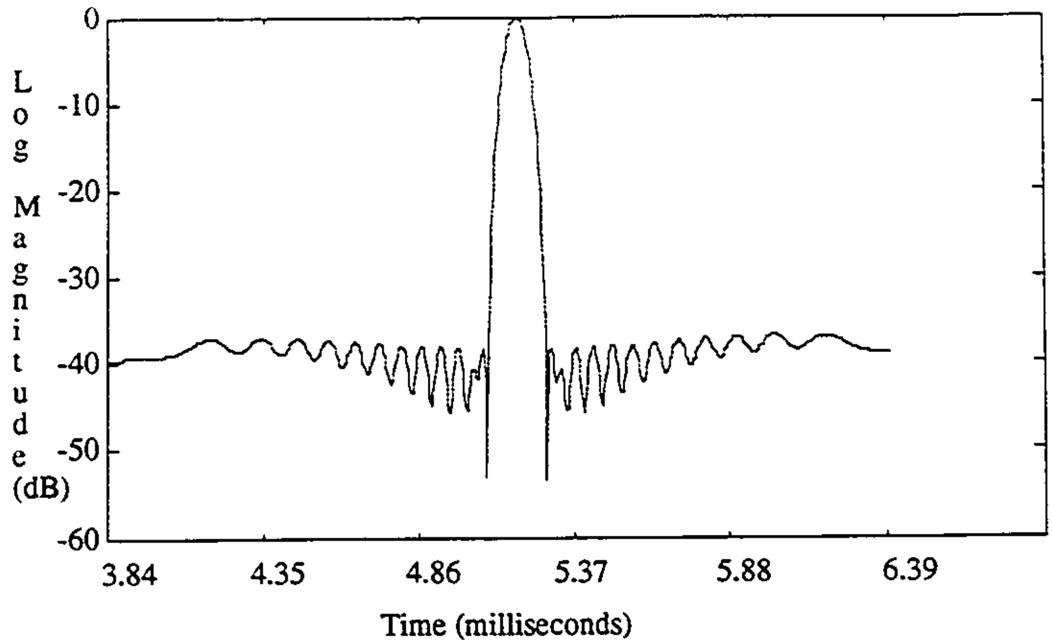

Figure 3.6(b). Log Magnitude plot of output Envelope of Filter with Hamming Weighting.

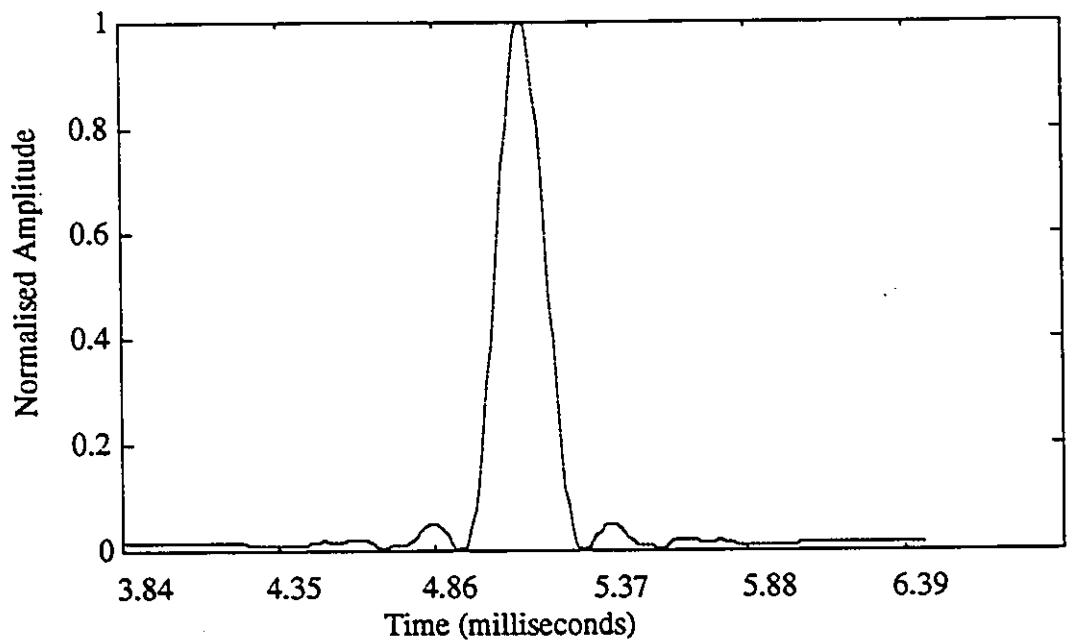

Fig 3.7(a) Output envelope of filter with Bartlett weighting.

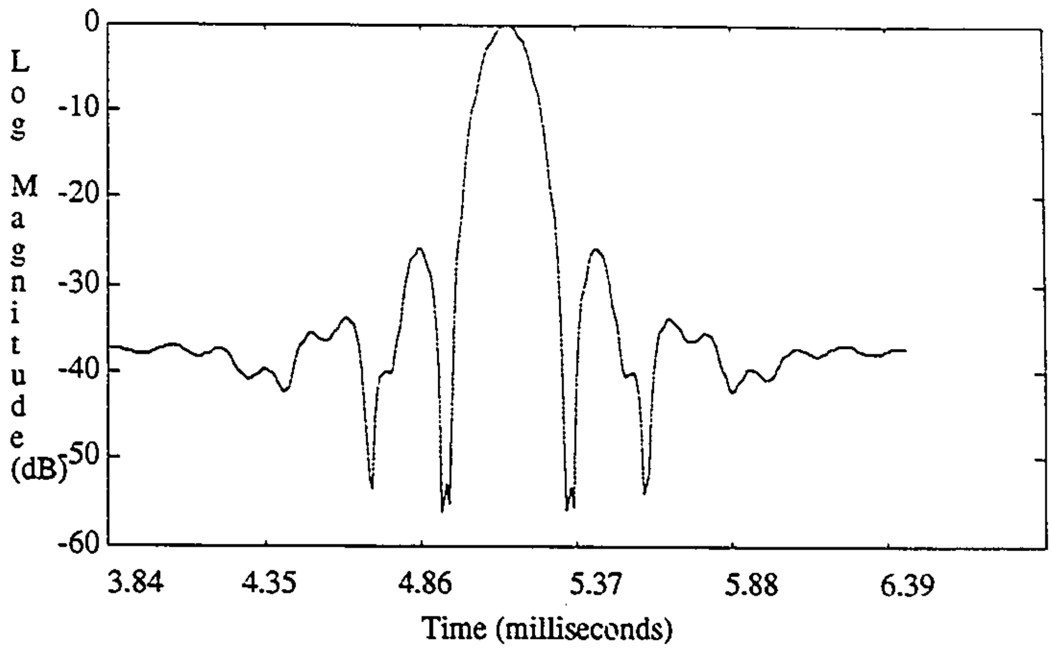

Figure 3.7(b) Log magnitude plot of output envelope with Bartlett weighting.

Window Type	3 dB Pulse Width (seconds)	Minimum Sidelobe Attenuation (dB)	Mainlobe Broadening
Unweighted	9.9650e-5	-13.6	1.0
Hamming	1.276e-4	-37.2	1.28
Bartlett	1.965e-4	-26.08	1.9719

Table 3.1. Comparison of filter output for different window functions.

3.6 Matched Filter Implementation

High speed convolution is a popular method of implementing matched filters. The main advantages of this method are the efficiency of the Fast Fourier Transform (FFT) used to perform the convolution and the flexibility it offers. Since signals emitted by the sonar is time limited, the operation of matched filtering is equivalent to filtering with a non recursive or Finite Impulse Response filter. The convolution operation is essentially equivalent to FIR filtering. The convolution of two discrete time signals $h(n)$ and $x(n)$. is given by:

$$y(n) = \sum_{k=-\infty}^{\infty} x(n)h(n-k) \qquad (3.16)$$

When a chirp signal of time duration T and of bandwidth Δ is sampled at n times its Nyquist rate it forms a discrete time signal of length $L = n\,\Delta\,T$. The time reversed version of this signal can be used as the impulse response of the matched filter.

Finite Impulse Response filters can be very efficiently implemented using high speed convolution. This method implements the convolution in the frequency domain. In this method two time domain signals of length N that need to be convolved are transformed into their Discrete Fourier Transforms. The two transforms are then combined by multiplying one with the other. The resulting product is inverse Discrete Fourier Transformed to yield the filter's output. This method involves the use of the Fast Fourier Transform and the Inverse Fast Fourier transform to compute the forward and inverse Discrete Fourier Transforms. For large values of N the computational savings over direct convolution in the time domain is very good. The disadvantage of this method however is that large data storage is required [Oppenheim]. On the other hand, direct convolution does not require the amount of data storage that high speed convolution needs but is computationally very expensive. Figure(3.8) shows the block diagram of a high speed convolver.

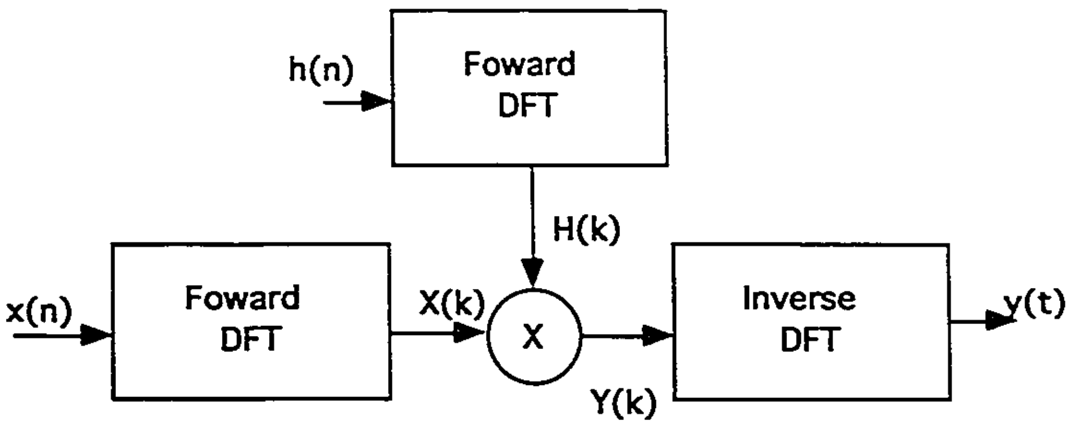

Figure 3.8. Block Diagram of High Speed Convolver.

Convolution performed using the Fourier transform is circular and each output batch of N points contains only $N-L+1$ valid output data points (N is the number of samples in the filter). In applications like ranging, continuous matched filtering is not needed. Rather, batch processing of all the data in one shot after the arrival of the echo would suffice. If M samples of data occur between the transmission of the pulse and its echo, then the size of the Fourier transform should be at least $N = M+L-1$. If N is not an integral power of 2, that is $N \neq 2^{v}$ where v is an integer, then N should be padded with zeros to arrive at a sequence whose length is an integral power of two.

The above does not preclude the manner in which the matched filter is implemented. Of course filters can be implemented in either hardware or software. Surface Acoustic Wave (SAW) devices find wide use in the implementation of matched filters, especially in radars. In the case of ultrasonic ranging these devices are not suitable as their operating frequencies would be in the Megahertz range, which is very much more than what the attenuation during propagation would permit. The size of these devices would be too large to operate in the ultrasonic range.

IV. Matched Filter Simulations

4.1 Introduction

The previous section on matched filters primarily considered only the detection problem, that is whether the echo from the target contained a replica of the signal(which may or may not have been buried in noise) or not. In the situation of range measurement not only does one have to detect the presence of signal in noise, but also has to determine the delay between the instant of transmission and arrival of the echo. In linear chirp systems the matched filter response should have a uniform amplitude and a time-frequency relationship that is of opposite slope to that of the transmitted wave form.

But the above is in conflict with the desirability of low time sidelobes. In making this tradeoff between low sidelobes and minimum signal to noise ratio degradation, the former is more important as high sidelobes can potentially mask a much weaker echo from a farther target. The reduction in the sidelobe level can be obtained by weighting of the amplitudes at the edges of the chirp pulse. This would result in the degradation of the signal to noise ratio of the matched filter(which is incidentally no longer truly matched to the signal), the amount of degradation depending on the type of weighting function chosen. Typically this mismatch loss is of the order of 1 to 2 decibels and this can only be compensated by increased transmitter power.

In the case of an ideal matched filter, the filter is basically the time reversed version of the signal it is matched to. In the case of pulsed ranging systems, a pulse of suitably modulated carrier is transmitted and the received echo is passed through the filter it is matched. This is essentially the convolution since the time reversed version of the signal acts as a window and moves past the echo. Each value of the matched filter output corresponds to one position of the window relative to the echo data.

The transmitted pulse can be thought of as a signal of infinite duration, truncated by a rectangular window, the duration of which is the same as the duration of the desired pulse. This is readily accomplished by multiplying the window with the infinite duration signal to obtain the desired pulse. The Fourier transform of a rectangular window function has the form of the sampling function, which is $\sin(x)/x$. If the signal were assumed to be a pure sinusoid, its Fourier transform would look like an impulse. Multiplication in the time domain is equivalent to convolution in the frequency domain. This means that a sampling function and an impulse are convolved in the frequency domain, thus leading to smearing in the frequency domain.

4.2 Spectrum Of Chirp Signals

For a linear chirp signal as the dispersion factor (also called the time bandwidth product) increases, the spectrum becomes more rectangular and thus most of the signal energy is concentrated in the band of frequencies of interest. Figures 4.1(a) and 4.1(b) shows the spectrum of a chirp signal for different dispersion factors. In both these cases the bandwidth during the generation of the chirp signal was kept a constant 10 kHz, while the time duration of the chirp pulse was changed. The time duration of the pulse in 4.1(a) was about 1 millisecond while for figure 4.1(b) was about 5.12 millisecond. Though they have the same frequency sweep, the spreading of the spectrum in the first case is obvious. This can be readily explained by the Fourier transform relationship.

Thus increasing the value of the dispersion confines the energy of the chirp to a bandwidth which approaches the swept bandwidth of the chirp $\Delta = kT$. From sampling theory, it is well known that for reconstruction of a discrete time signal the original continuous time signal has to be sampled at a minimum of twice the highest frequency component. However this is true only in the lowpass case, that is the lowest frequency component in the signal is zero. For the case of bandpass signals however, the continuous time signal needs to be sampled

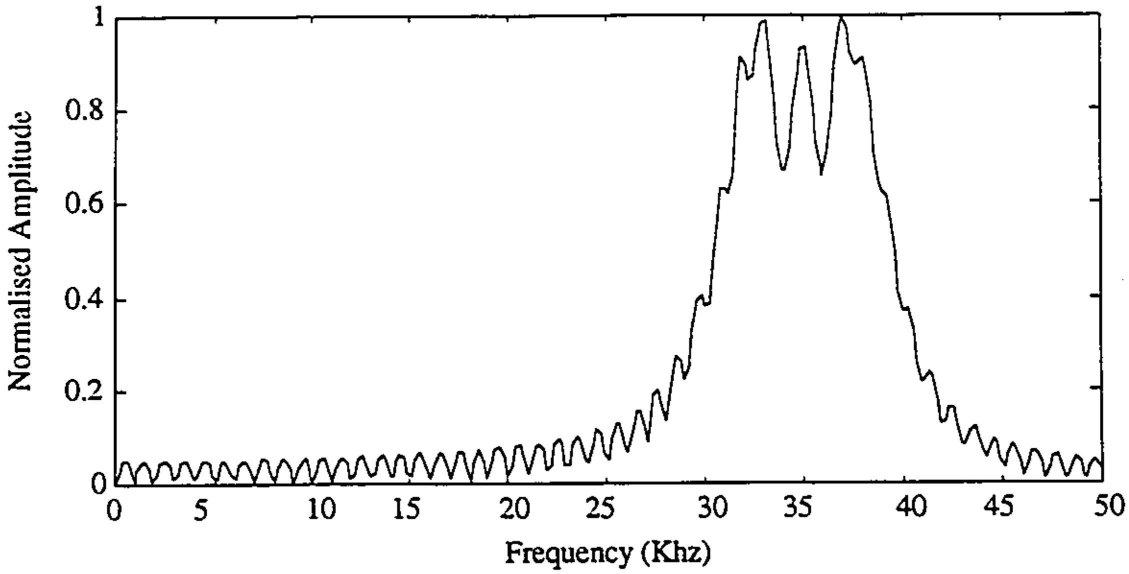

Figure 4.1(a).Spectrum For a Chirp with Dispersion Factor = 10.0.

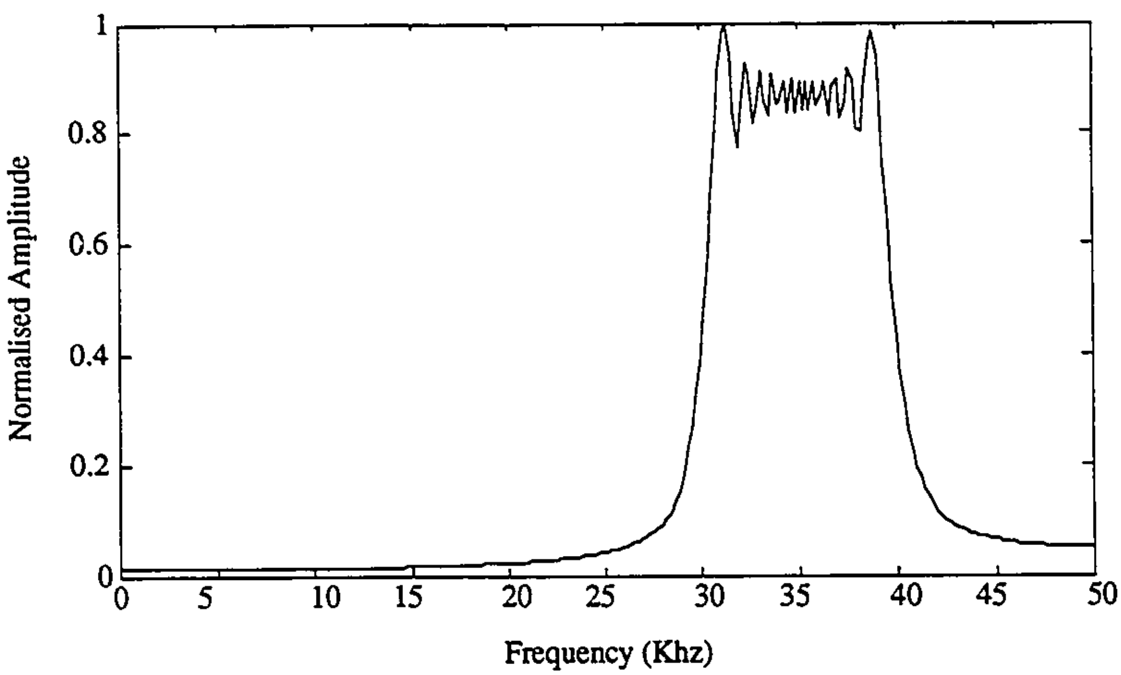

Figure 4.1(b) Spectrum For a Chirp With Dispersion Factor =52.22.

at a rate no higher than twice the bandwidth of the signal. The original signal can be reconstructed from this sampled signal by passing the sampled signal through a bandpass filter with cutoff frequencies corresponding to the edges of the band of the signal[Taub, Schilling]. This is of particular importance especially in microwave radars since their operating frequencies are in the order of several hundred MHz and well into the GHz range. This can significantly reduce the amount of data that needs to be processed.

4.3 Amplitude Weighting to Reduce Sidelobes.

As discussed earlier, the output of an unweighted matched filter for a chirp signal has a minimum sidelobe level of about 13.5 dB below the main lobe peak. The high levels of the sidelobes can potentially mask a much weaker echo near it. Thus the side lobe levels have to be reduced in order to increase the detectability of weaker signals especially in the presence of strong echoes. One method of achieving this is through amplitude weighting of the frequency components in the chirp signal.

As mentioned earlier, windows find wide use in FIR filter design. This section deals with the application of window functions to reduce the sidelobes produced by an uniformly weighted matched filter. The window functions applied perform the task of amplitude weighting the frequencies in the chirp signal. When this is imposed equally on the transmitting and receiving stages the resulting filter is still matched. However, in most applications this is done at the receiver alone. The main reason for this being that at the transmitter the power transmitted is more important and in most radar applications the amplitude weighting at the transmitter would mean inefficient use of the transmitting tubes. The same can be said of sonars. Thus in most cases the amplitude weighting is performed at the receiver alone, resulting in a filter that is now no longer exactly matched to the transmitted signal.

The frequency response of a matched filter is adjusted to maximize the output signal to noise ratio and requires that the amplitude response of the matched filter be equal to the amplitude spectrum of the input signal. The amount of mismatch is measured as a degradation from the optimum signal to noise ratio normally produced with a matched filter. When the weighting is used in the receiver alone the, the loss in signal to noise ratio is characterized by the loss factor L_s , which is given by:

$$L_s = \frac{(S/N)_{weighted}}{(S/N)_{unweighted}} \tag{4.1}$$

The amplitude weighting function chosen should provide minimum mainlobe widening for a specified sidelobe level. The amplitude weighting functions that will lower the sidelobes in the output time waveform all have the characteristic bell or tapered shape. When the window is applied to the chirp, the center frequency (the carrier of frequency f_0) will have the maximum amplitude, while all other frequencies will have decreasing amplitudes about the center frequency.

The commonly used window functions are defined in the following equations. All the windows are considered to of length N.

Rectangular:

$$w(n) = 1, 0 \le n \le N-1 \tag{4.2a}$$

Bartlett:

$$w(n) = \begin{matrix} \dfrac{2n}{N-1}, 0 \le n \le (N-1)/2 \\ 2 - \dfrac{2n}{N-1}, (N-1)/2 \le n \le N-1 \end{matrix} \tag{4.2b}$$

Hanning $(\alpha = 0.5)$ and Hamming $(\alpha = 0.54)$:

$$w(n) = \alpha + (1-\alpha)\cos\left(\frac{2\pi n}{N-1}\right), 0 \le n \le N-1 \tag{4.2c}$$

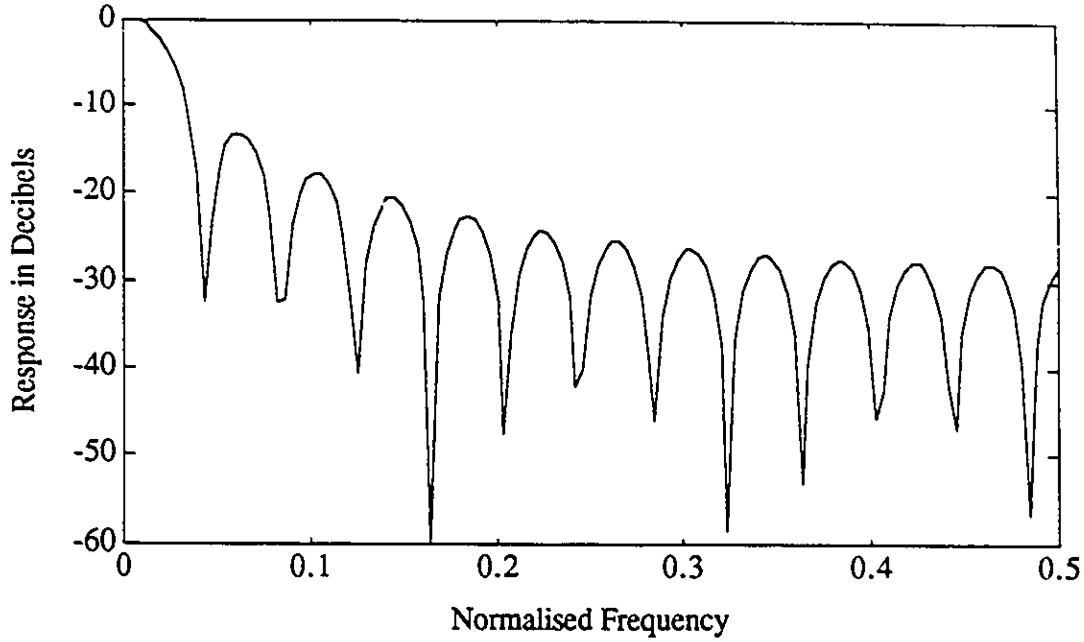

Figure 4.2a. The Fourier Transform Of a Rectangular Window.

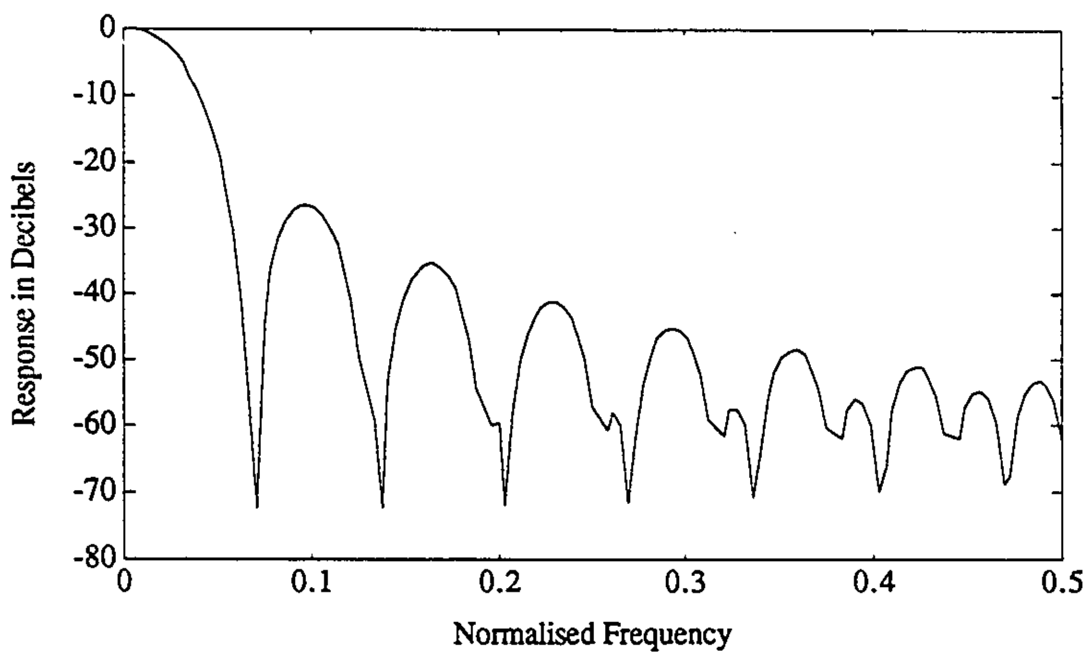

Figure 4.2b. The Fourier Transform of a Bartlett window.

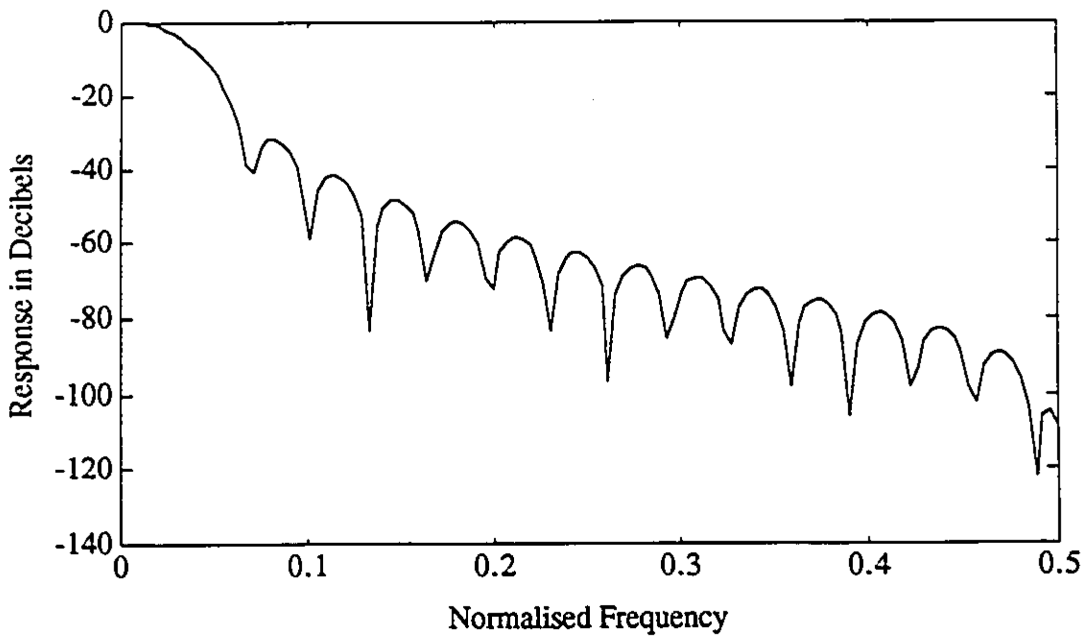

Figure 4.2c. Fourier Transform of a Hanning Window.

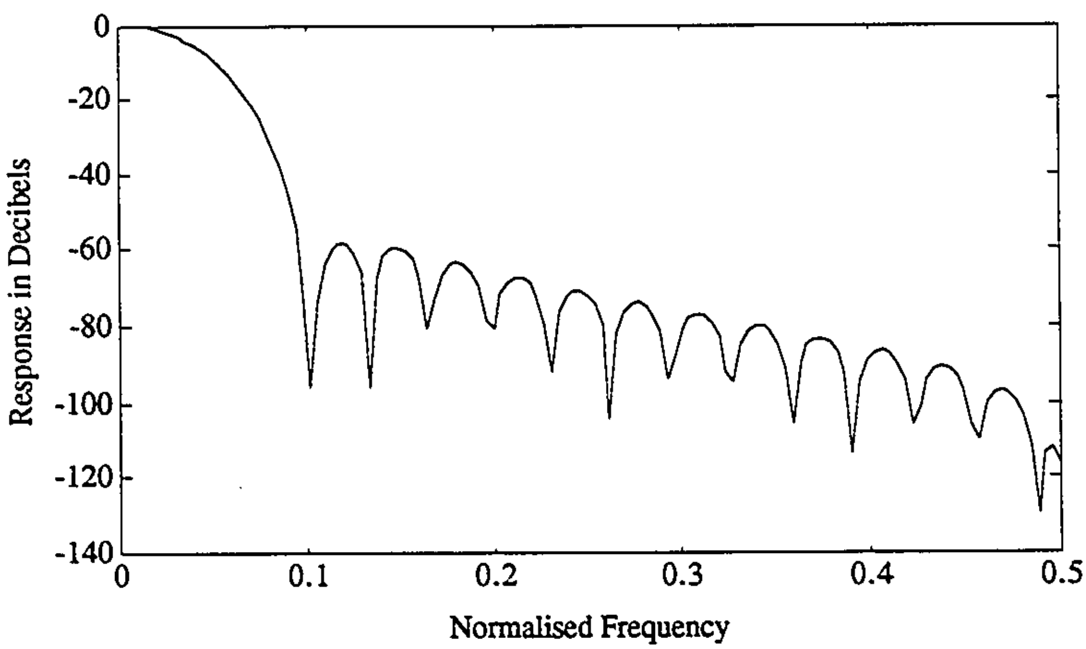

Figure 4.2d. Fourier Transform of a Blackman window.

Blackman:

$$w(n) = 0.42 - 0.5\cos\left(\frac{2\pi n}{N-1}\right) + \cos\left(\frac{4\pi n}{N-1}\right), 0 \le n \le N-1 \quad (4.2d)$$

Since all the above windows are symmetrical the phase is linear. The rectangular window has the narrowest main, but the side lobe levels are relatively high. Figures 4.2a through 4.2d show the Fourier Transforms of the windows defined in equations 4.2a-d.

The process of amplitude weighting the coefficients of the matched filter is quite simple to implement. First a suitable type of window is to be chosen. The length of the window should be so determined so as to be the same as that of the chirp pulse. The structure of the chirp is already known to the designer. The window and a time reversed version of the chirp should be multiplied element wise in order to obtain the amplitude weighted matched filter for the chirp. The frequency response of a hamming weighted matched filter for a chirp with center frequency 35.110 kHz, swept bandwidth 10.22 kHz, duration 5.12 milliseconds, and dispersion factor 52.2 is shown in Figure 4.3a. The Frequency response of the actual matched filter is also shown. The output of the Hamming weighted matched filter when the original chirp is passed through it is shown in figure 4.3b. A single chirp pulse was passed through the matched filter with no delay with respect to the instant the filter was activated. The filter length in this case was 512 corresponding to the number of samples in the chirp pulse. Thus the first 512 output samples of the matched filter correspond to time aliased data and has to be discarded. The peak in this case occurs at the 513th sample corresponding to zero delay.

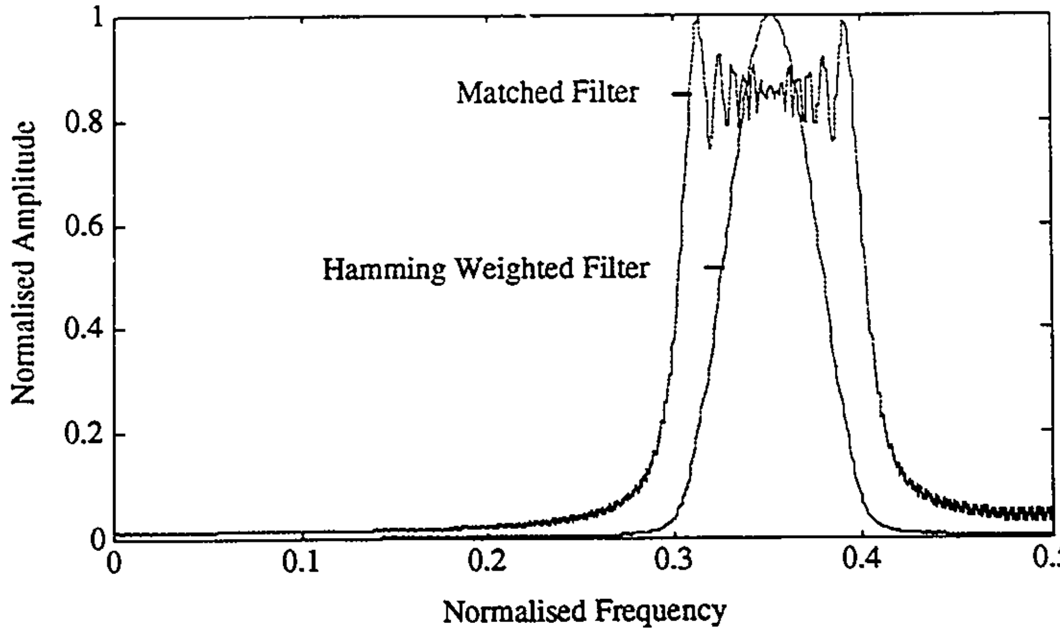

Figure 4.3a. Frequency response of Matched Filter and Hamming weighted filter.

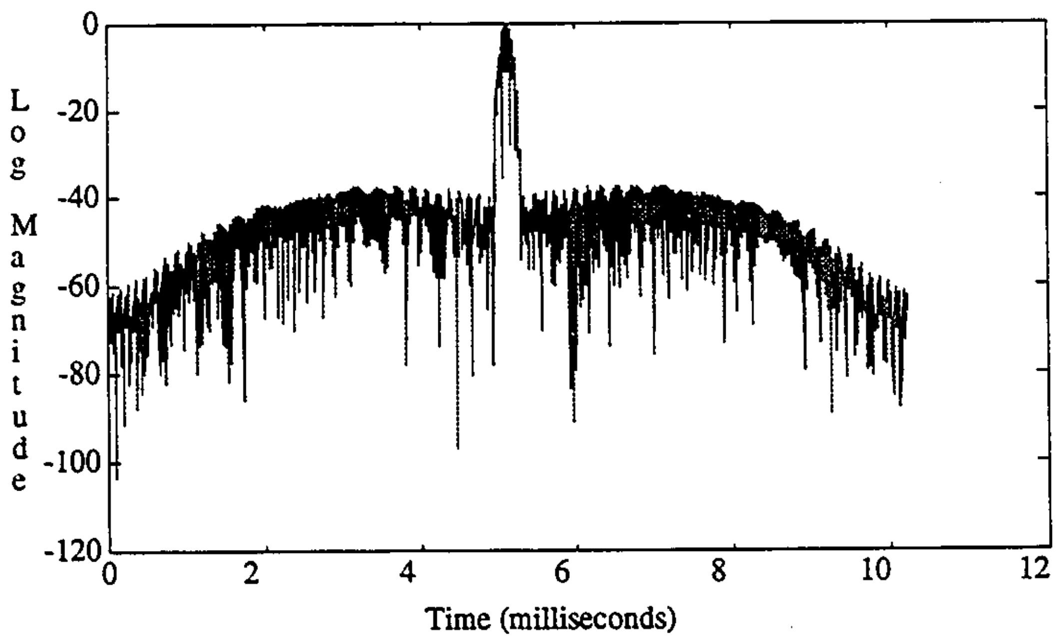

Figure 4.3b. Log Magnitude Plot of Output of Matched Filter With Hamming Weighting.

4.4 Range Determination.

Range estimation is essentially the determination of time delay. If the amount of delay from the instant a pulse is transmitted to the instant it is received back is known, the range R to the target can be computed as :

$$R = \frac{ct}{2} \tag{4.3}$$

where t is the total time delay for the signal to reach the target and back and c is the velocity of propagation in the medium the signal transmitted in. The maximum range that can be detected by a typical pulsed system depends on the power transmitted by the transmitting transducer. This is given by :

$$P_r = \frac{KP_{Tr}}{R^4} \tag{4.4a}$$

Or $$R = \sqrt[4]{\frac{KP_{Tr}}{P_r}} \tag{4.4b}$$

where P_{Tr} is the peak transmitted power from the transducer, P_r is the peak received power at the transducer and K is a constant of proportionality that accounts for transducer gains, target cross section, attenuation during propagation and the like. Thus it is obvious from equations (4.4a) and (4.4b) that the received power at the transducer is inversely proportional to the fourth power of the range. In order to increase the range two fold the transmitted power has to be increased by a factor of sixteen.

For a Chirp signal defined by equation 3.1b with center frequency 35.11 kHz, bandwidth of 10.22 kHz, duration 5.12 ms and sampled at 100 kHz (total of 512 samples), the energy contained in the signal is given by

$$\sum_{n=0}^{N-1} |S(nt)|^2, 1 \leq N \leq 512 \tag{4.6}$$

For this particular case the energy (across a 1 ohm resistance) in the signal computes to be 256.7 joules. Now in the absence of noise if this signal were passed through its own matched filter the peak amplitude of the output pulse should equal this amount. Figure 4.4 shows the output of the matched filter matched to this particular signal. No amplitude weighting was employed in this case. As can be seen from Figure 4.4(a) the peak occurs at sample number 513 corresponding to zero time delay with an amplitude of 256.68. This is in agreement with the calculated value of the output peak. Only 128 samples of the output are shown in this plot for clarity. The swept bandwidth (Δ) of the signal is 10.22 kHz. In the case of an unweighted matched filter, the 3 dB width of the compressed pulse (mainlobe) is equal to $1/\Delta$. The spacing between the first zeros in this case should be equal to :

$$\frac{1}{\Delta} = \frac{1}{10220} = 97.947\mu s \text{seconds}.$$

The compressed pulse in the output of the unweighted matched filter has a width of 20 samples corresponding to $99.5\mu s$ seconds. The minimum sidelobe level in this case is 13.6 decibels below the main lobe peak.

For the Hamming weighted matched filter in figure 4.4b, the 3 dB width of the compressed pulse is $127\mu s$ seconds. The minimum sidelobe level is 37.2 dB below the peak of the mainlobe. The width of the compressed pulse for the filter with Hamming weighting is 1.28 times wider than that for the unweighted filter.

At time t=0 a chirp pulse of time duration T is transmitted by the transducer. The same signal is transmitted (though this need not always be the case) after a duration t_r , where t_r is called the inter pulse period and $f_r = 1/t_r$ is called the pulse repetition frequency. When the same transducer is used for both transmission and reception of signals, it is not possible to start receiving the echoes until time t=T, that is the transducer can go into receive mode only after the transmission has been completed. The potential reception interval extends from t=T to t=t_r and contains the **range** and **reception** window. If the target of interest lies in a range interval bounded by R_{min} and R_{max}, the **range window** is defined as the

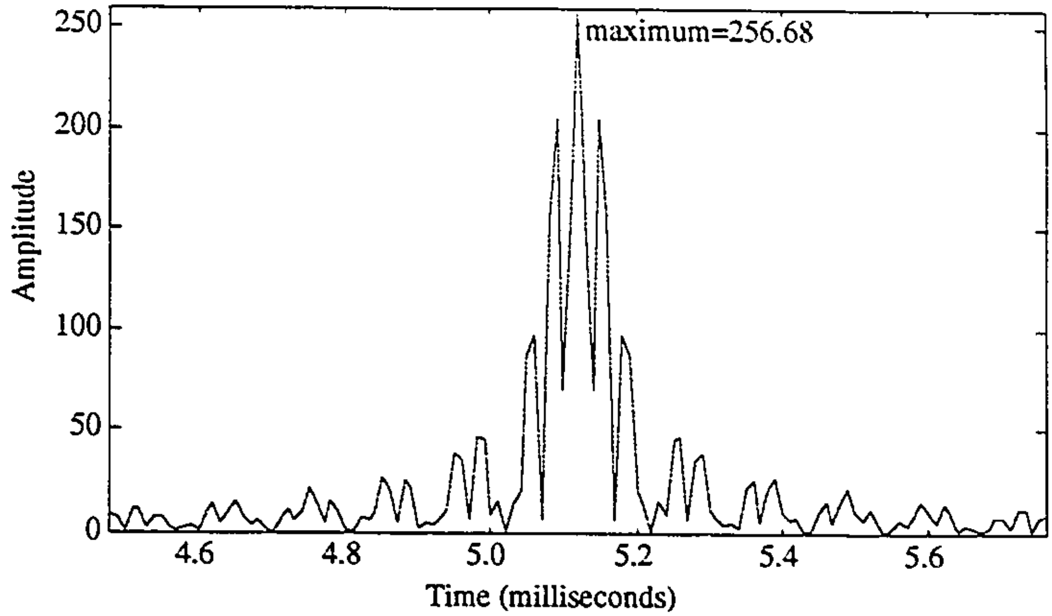

Figure 4.4(a) Output of Unweighted Matched Filter For Signal Described Above.

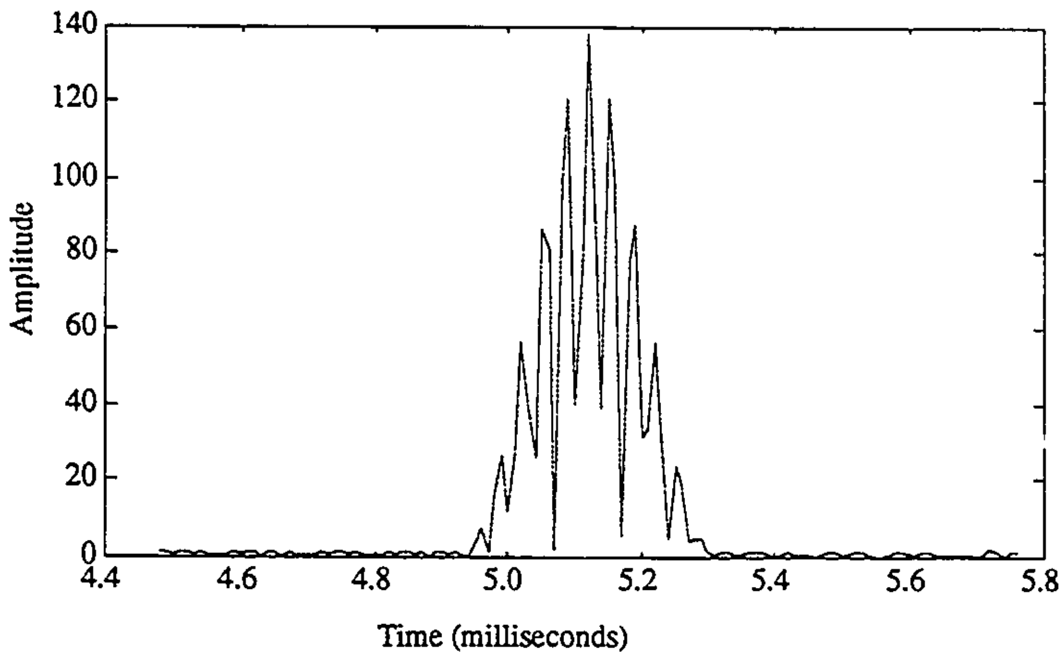

Figure 4.4b. Output of Matched Filter With Hamming Weighting.

interval between $t_1 = 2R_{min}/c$ and $t_2 = 2R_{max}/c$, during which echoes return from the targets. Since pulse compression has to be employed in the case of Chirp ranging systems (though this is not unique to Chirp ranging systems only), it is necessary to receive data for a time duration equal to the range window plus the signal length T. This duration is called the **receive window.** This parameter is important in systems as it gives an indication as to how much data has to be processed.

4.4.1 Noise Equivalent Bandwidth

In the preceding sections the basic assumption made was that only the signal was present without any additive noise. In the presence of noise, when the received echo is passed through the matched filter, the output of the matched filter is no longer the same as in the absence of noise.

The input to the matched filter consists of the return echo plus white noise. The power spectral density of white noise is given by :

$$G_n(\omega) = \frac{\eta}{2} \text{ for all } \omega \qquad (4.7)$$

This means that the power spectrum is constant at all frequencies. This describes the two sided power spectrum. Though white noise has infinite bandwidth, the chirp signal and its corresponding matched filter have a finite bandwidth. As a result the filter does not pass all the input noise to the output, but only a certain band of the interfering noise. A parameter that is frequently used to characterize the bandlimiting characteristics of a filter is **noise equivalent bandwidth B_n** [Stremler]. This is defined as that ideal filter of bandwidth B_n that would pass the same amount of noise power as the system. The power density at the output of this filter is white within the bandwidth B_n and zero everywhere else. Also the area under this rectangular power spectral density is equal to the area under the spectral density of the actual filter. This is illustrated in Figure 4.5. The dotted line in the figure indicates the actual noise components passed by the filter, and the solid line the noise passed by an ideal filter. For a signal $s(t)$ the noise bandwidth is given by:

$$B_n = \frac{\int_0^{\infty} |F(\omega)|^2 df}{\max(|F(\omega)|^2}\tag{4.8}$$

where $F(\omega)$ is the Fourier transform of $s(t)$

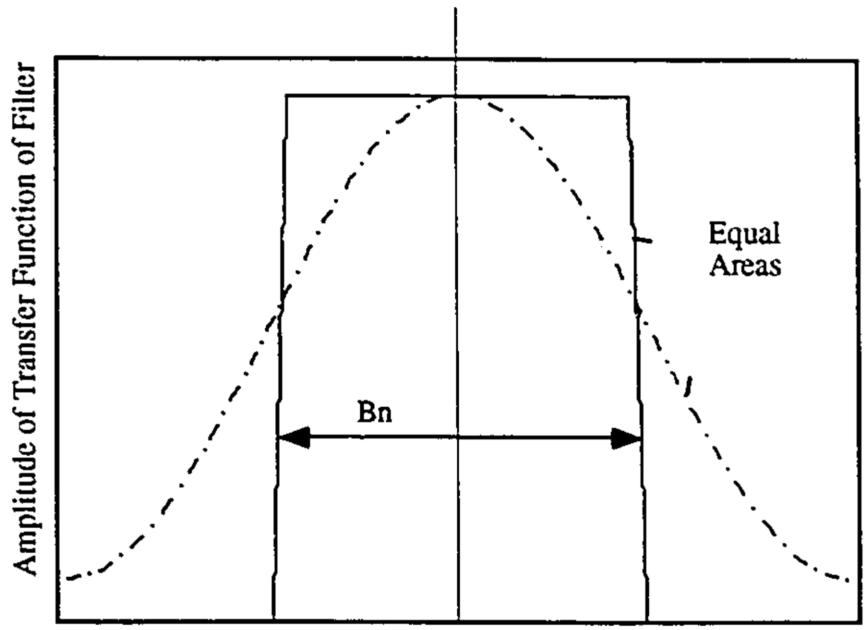

Figure 4.5. Noise Bandwidth of a Filter

As mentioned earlier, the matched filter is used for detection of signals buried in white gaussian noise and not for signal fidelity. Figure 4.6a shows the output of an unweighted matched filter, for an input signal to noise ratio -3.3 dB. The corresponding output signal to noise ratio was 16.46 dB. The pronounced peak indicated the detection of the signal buried in noise. Figure 45.6b shows the output of a filter that was amplitude weighted with a hamming window. Note that the relative magnitude of the peak is lower than for an

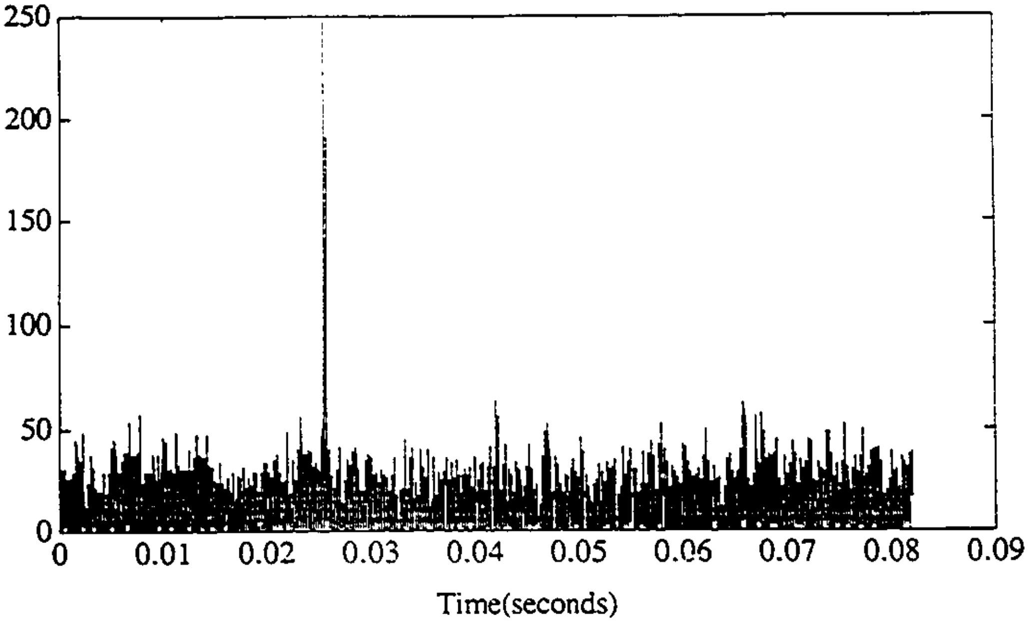

Figure 4.6a Matched filter output for Signal with noise present.

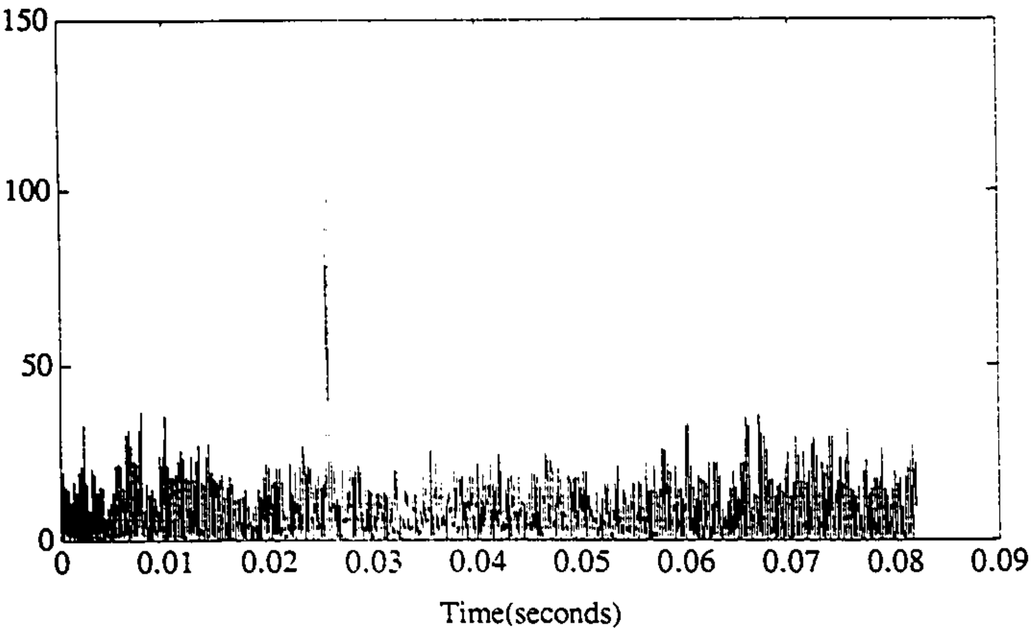

Figure 4.6(b) Hamming Weighted Matched filter output with noise present.

unweighted matched filter.

4.5 The Simulation Environment.

All the above simulations were performed on a Banshee system installed on a 486 machine. The Banshee system is a full length AT compatible board that contains a TMS320C30 processor that can perform 33 million floating point operations per second[ASPI]. The software used for the above was SPOX which is a Digital Signal Processing (DSP) operating system that can meet the needs of high end DSP processors like the TMS320C30. The advantage of using SPOX is that the operation of the Banshee board is transparent to the user. It offers the user to go about developing DSP algorithms without getting involved with the details of operation of the processor. The simulations were also tested using MATLAB.

V. The Polaroid Ranging System

5.1 Introduction

As supplied by the manufacturer, the Polaroid ultrasonic ranging system consists of two major elements, which are the transducer and the ranging module. During operation of the system a pulse is transmitted towards a target and the resulting echo is detected. The elapsed time between the initial transmission and the echo detection is then converted to distance with respect to the velocity of sound [Polaroid].

The main component in this system is the transducer which acts as both the loud-speaker and the receiver. When the unit is activated the transducer emits a sound pulse and then waits for the echo from whatever target it is aimed at. Figure 5.1 shows the block diagram of the system. The ranging module consists of two integrated circuits, one analog and one digital. In its default configuration the system has a maximum range of about 35 feet. In addition to the ranging module, also available from Polaroid is the Experimental Demonstration Board (EDB). By connecting the ranging module to the EDB, ranging can be performed to a maximum distance of 35 feet. The EDB provides a LED display of the range information.

The Polaroid transducer was first used for automatic focusing of a range of cameras manufactured by Polaroid. In this particular application the transducer excelled mainly because of its high reliability in low light level situations. A thin foil is the moving part which transforms electrical energy into ultrasonic waves and vice versa. The foil is plastic (Kapton) with a conductive gold coating on the front side. This is stretched over an Aluminum backplate. The backplate and foil represent a capacitor. When charged an electrostatic force is exerted on the foil. An Alternating Current (AC) voltage of given frequency forces the foil to vibrated at the same frequency and to send out ultrasonic

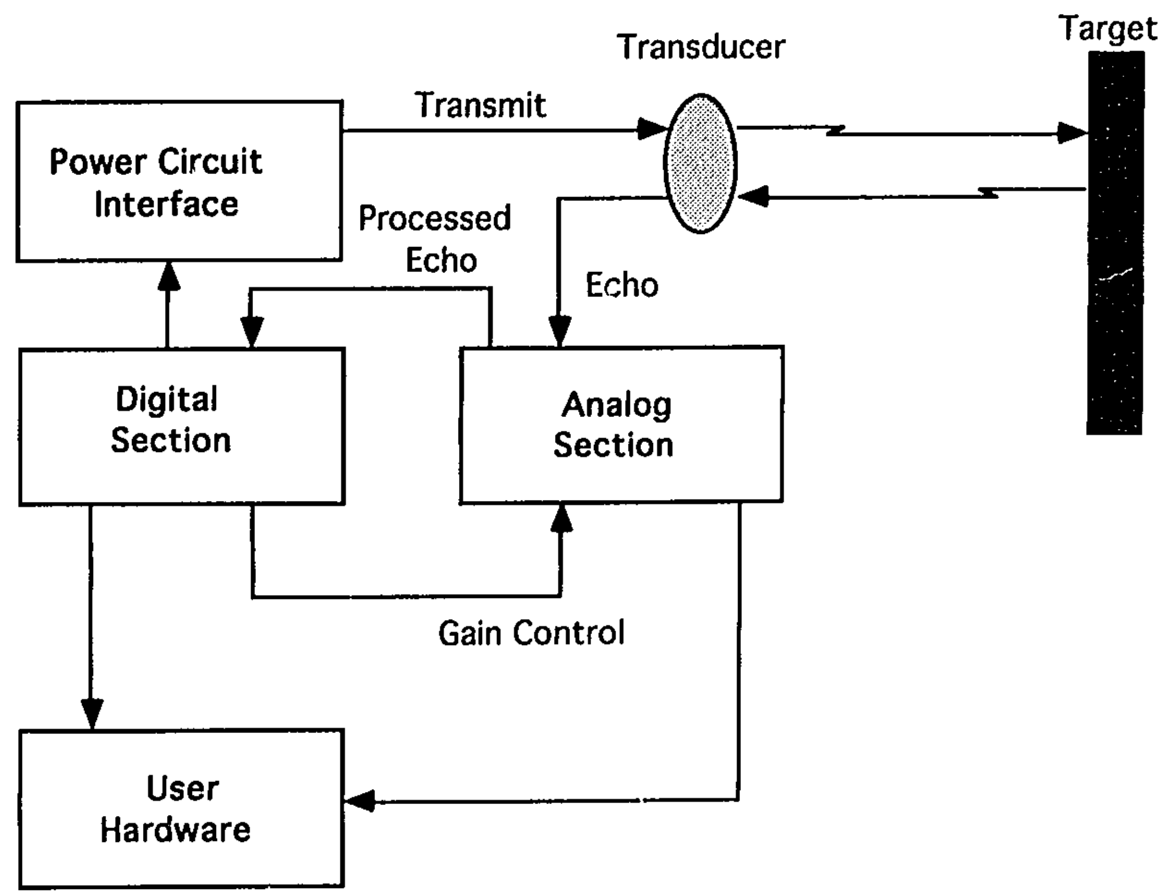

Figure 5.1. Block Diagram of the Polaroid Ranging System.

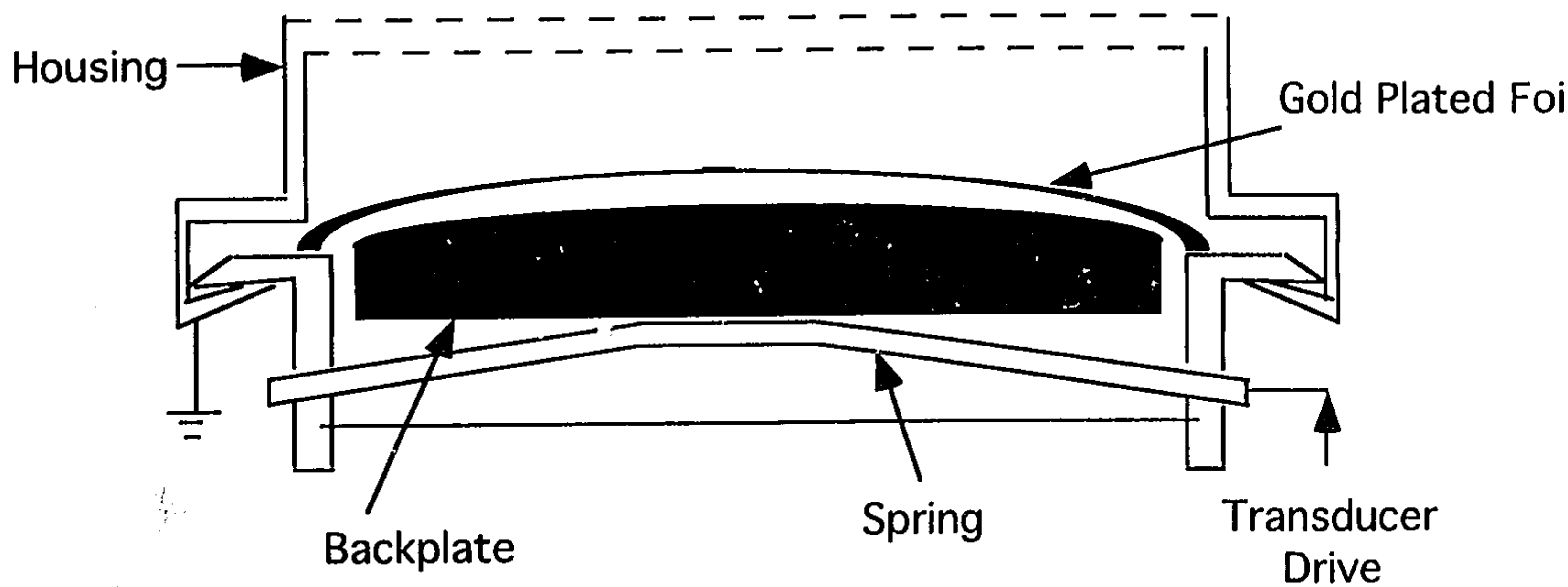

Figure 5.2 The Transducer Assembley

waves. A steel spring maintains contact with the backplate and holds the foil under constant tension. A perforated front cover mechanically protects the foil. Figure 5.2 shows the transducer assembly.

In any ranging application the received echo is definitely going to be very much weaker than the transmitted signal. Thus the received echo has to be amplified . However this amplification should not be constant, rather the amplification should be varied with distance. It is intuitive that the received echo from a target close to the transmitter would be stronger than a echo from a target far from it. Its therefore desirable to keep the amplification of echoes from nearby targets low, and those from farther targets high. Since the round trip time of the signal is proportional to the distance, the amplification should increase as a function of time. The ranging module incorporates an amplifier in the analog IC, whose gain increases with time.

5.2 The Ranging Module

This section describes the operation of the ranging system as supplied by the manufacturer. The two integrated circuits on the ranging board are the SN28784 which is the analog IC and TL851 which is the digital IC. The schematic of the ranging board is given in figure 5.3. When pin 14 (INIT) of the TL851 goes high, the chip internally generates 16 pulses of frequency 49.5 kHz and this drives the base of the transistor. This produces the drive for the transducer. At the end of the 16 pulses a DC bias of 200 volts will remain on the transducer as recommended for operation. In order to eliminate detection of ringing of the transducer as a return signal, the receive (REC) input of the TL851 is inhibited for 2.38 milliseconds after INIT goes high. If a reduced blanking time (in applications where distances less than 35 centimeters need to be measured), BINH input of the TL851 can be taken high as long as the transducer damping is sufficient so that ringing is not detected as the return signal.

During a cycle starting with INIT going high, the receiver amplifier (in SN28784) gain is increased in discrete steps, since the transmitted signal is attenuated with distance. This is done by the four gain control pins (GCA through GCD) in the TL851. The maximum gain can be increased by increasing the value of resistor R1. However there is an inherent trade off in this case as increasing the receiver gain while enhancing the detectability of the echo also increases the susceptibility to noise. The echo from the target is amplified and appears as a high logic level output at the ECHO pin of the TL851. The elapsed time between the INIT pin going high and ECHO going high is proportional to the distance between the target and the transducer. If the distance to multiple targets (i.e. more than one target lies in the path of the ultrasonic beam) need to be measured, then the cycle is slightly different. After the first return signal that causes the ECHO output to go high, the blanking (BLNK) input must be taken high and then back to low in order to reset the ECHO output [Polaroid].

5.3 User Interface

The previous sections described the ranging module if operated in conjunction with the EDB. During the entire ranging operation INIT should be high. INIT going low resets the ranging module. The INIT signal is supplied by the EDB and it remains high for 100 milliseconds before going low. In order to extend the range first and foremost INIT has to be kept high for longer than 100 ms. Also the TL851 generates pulses of a constant 49.5 kHz. Thus in order to extend the range of the system by transmitting a chirp signal, INIT has to be kept high for at least 120 milliseconds corresponding to a round-trip distance of about 40 meters or range of 20 meters.

The Banshee board as introduced in the previous chapter in addition to the TMS320C30 processor also contains an Analog to Digital (A/D) and Digital to Analog (D/A) daughter board. This board is a 16-bit system that operates at up to 100000 samples per second sampling rate. It contains two 16-bit A/D and two 16-bit D/A converters. It thus allows for easy Input-Output (I/O) interface especially when used in conjunction with SPOX.

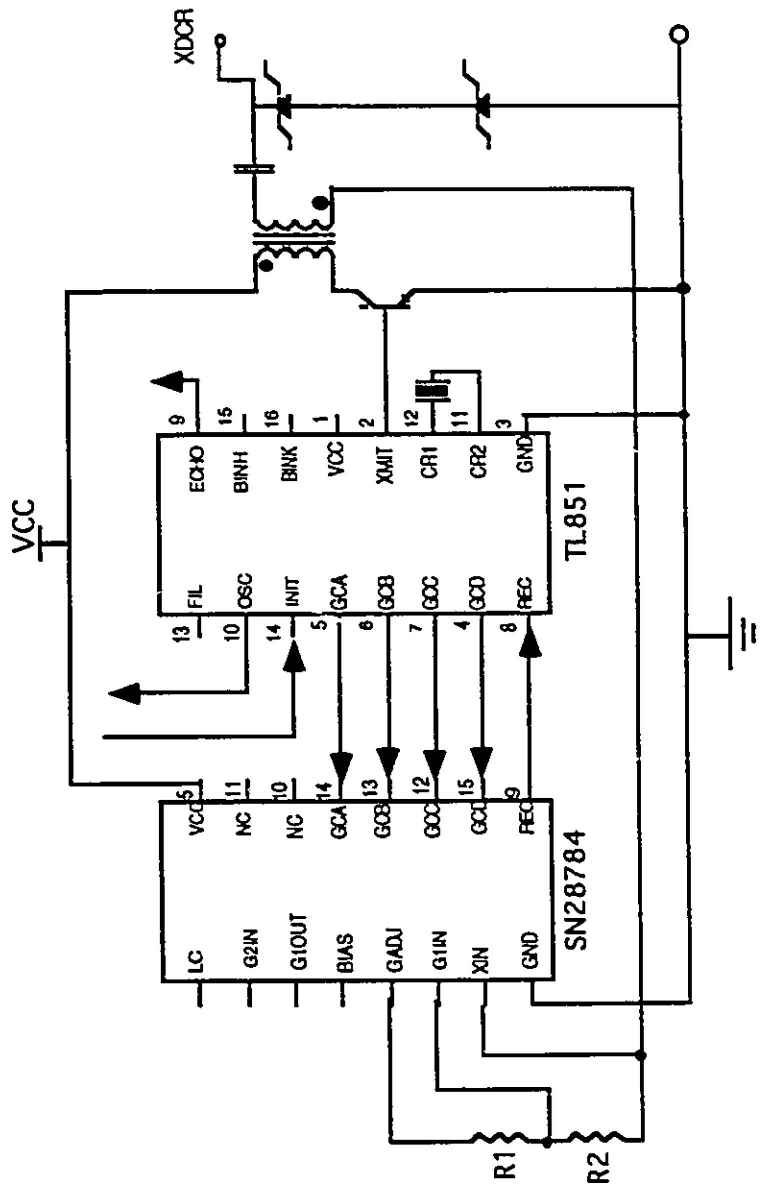

Figure 5.3. Schematic Of the Ranging Module

SPOX has a built in streaming I/O interface. A stream is a channel through which data flows between an application program and an I/O device. The advantage of using this is that it provides a simple and universal interface to all I/O devices, hiding the details of the device operation [SPOX].

Figure 5.5 shows the flow diagram of the ranging operation. There are three distinct I/O operations that are involved. The first is to send the INIT input to the ranging module. The second is to send out the CHIRP pulses to drive the transducer (transmit mode). The third is to acquire the echo from the ranging module (receive mode). Thus there are two output and one input operations.

The INIT input has to be kept high for as long as the ranging is in progress. This can be accomplished by connecting INIT to one of the D/A channels of the Banshee daughter board and keep the channel at a high level for the duration of the ranging. This however would prove troublesome as the other I/O channels cannot be used concurrently. In order to overcome this problem a very simple interface circuit was built as shown in Figure 5.4. This was implemented using the 7474 dual D-type positive edge triggered flip-flop. The

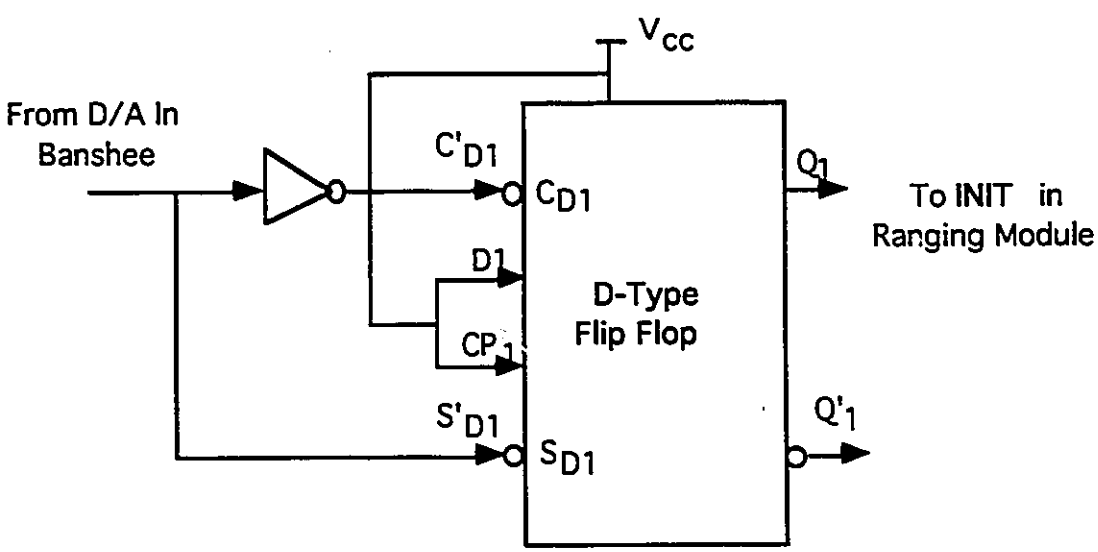

Figure 5.4 Schematic of Interface to keep INIT high for the duration of Ranging.

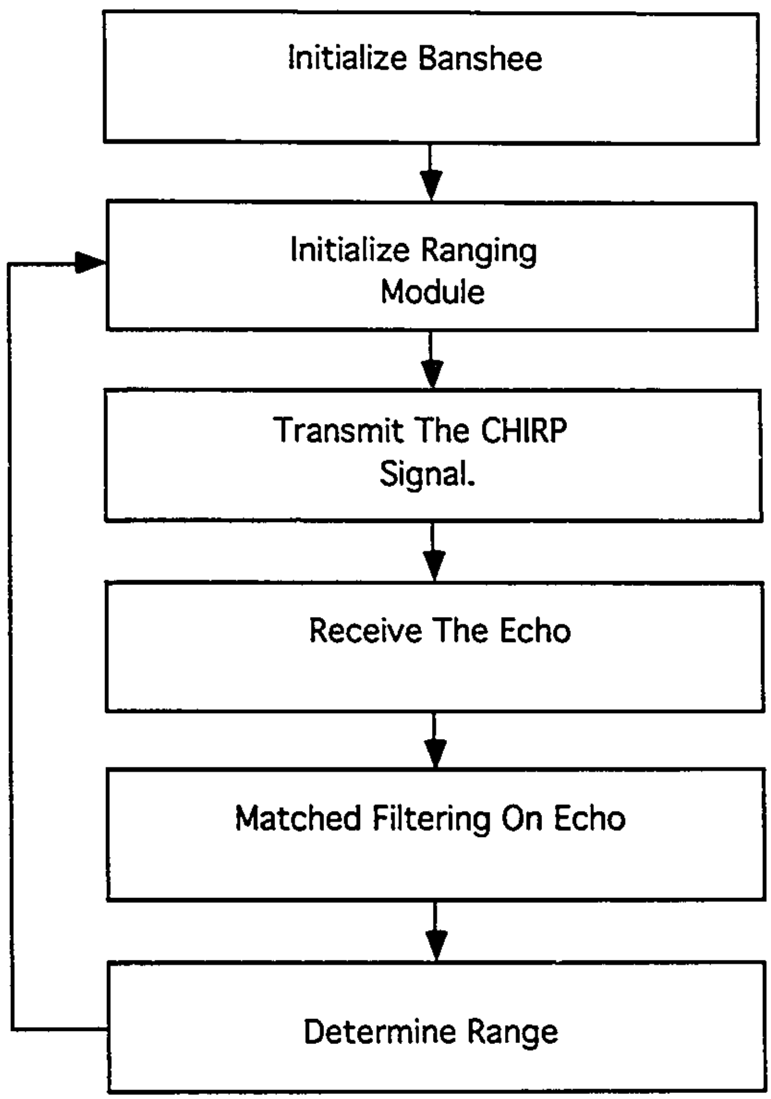

Figure 5.5 Flow Diagram Of Ranging Operation.

Clear (C'_{D1}) and **Set** (S'_{D1}) inputs are active low. The information at the input (D_1) is transferred to the output on the positive edge of the clock pulse (CP_1). Complementary outputs are available, Q and Q'. After the clock pulse input threshold is passed the input data is locked out and information will not be transferred to the outputs until the next rising edge of the clock pulse. **Set** and **Clear** are independent of the clock, that is regardless of the state of the clock pulse if Set is made low Clear high, the output Q would be high and vice versa. It is this property of the flip flop that is taken advantage of in this interface circuit. The input and clock pins are tied to the power supply. The output of the D/A converter from the Banshee is connected to the Set input and it complement to the Clear input. The output Q1 is connected to the INIT pin in the ranging module. When the ranging module has to be initialized, the output D/A channel is taken high briefly. This would cause the flip-flop to be cleared, that is the data at pin Q1 would be low. Once this is done, the D/A channel can be taken low, in response to which the output Q1 would go high. So long as that particular D/A channel is not used for any other I/O, it would continue to maintain a low level (close to zero volts). As a result the output pin Q1 would remain high until such a time when the D/A is taken high . Thus one D/A channel is dedicated for the purpose of keeping INIT high.

The next step is to transmit the chirp signal. The drive signal should be fed to the base of the NPN transistor in the ranging module. The signal should be in the form of pulses of the desired frequency to be transmitted. The signal can be generated in software and output to the Ranging module through the other D/A channel on the Banshee board.

The process of receiving the data is to open one A/D channel and sampling the amplified echo at pin 6 (G1OUT) of the SN28784 on the Ranging module. Since the maximum range to be measured is about 20 meters, it is sufficient to sample G1OUT for about 120 milliseconds. In all the cases the frequency at which the echo was sampled was 100 kHz. This constitutes the data acquisition phase. Once the echo data has been acquired, the next step is to pass the received data through its own matched filter, which is easily implemented in software as discussed in chapters 3 and 4. The output of the matched filter would be a stream of numbers and the location of the peak can be readily determined. Since the sampling frequency is known, the location of the peak in the output of the matched filter gives the delay between the transmitted and received signals. Once the time delay is known it is easy to calculate the distance to the target as the velocity of propagation of sound is known.

5.4 Results

The Polaroid ultrasonic transducer has a flat frequency response from 20 kHz to 100 kHz. This implies that the swept bandwidth in the transmitted chirp pulse can be in the tens of kilohertz. Thus for even pulses of short duration, large values for the dispersion factor can be obtained. For example, for a pulse duration (T) of 1 millisecond and a swept bandwidth (Δ) of 50 kHz the value of the dispersion factor (D) would be 50.

As discussed earlier in section 4.2 the minimum sampling frequency when sampling a continuous time signal should be at least twice the highest frequency component present in the continuous time signal. Normal practice is to sample signals at frequencies higher than this minimum. The Banshee board that was used for data acquisition can sample data at a maximum frequency of 100 kHz. This condition limited the highest frequency component present in the chirp pulse to 50 kHz. In section 4.2 it was shown that the chirp pulse could have considerable frequency content outside of its swept bandwidth depending on the value of the dispersion factor. This dictates that the highest frequency that should be used in the chirp signal should be well below 50 kHz in order to prevent aliasing.

Ranging using the procedure discussed in the previous section was performed. The target in all cases was a wall at varying distances from the transducer. The following describe the transmitted chirp pulse and other system parameters:

Center Frequency273°of the chirp pulse f_0	= 35.11 kHz
Duration of the chirp Pulse T	= 1.28 ms
Swept Bandwidth of the chirp pulse Δ	= 7.812 kHz
Dispersion Factor of the chirp pulse D	= 10.0
Sampling frequency f_s	= 100 kHz

The Fourier Transform of the transmitted pulse is shown in figure 5.6. Tables 5.1 through 5.3 show the results of the ranging operation for target distances of 10.2 meters, 12.3 meters and 13.5 meters respectively. The measurements are rounded to the nearest centimeter. The temperature during the ranging operation was 22° Celsius or 295° Kelvin. At 0° Celsius (273° Kelvin) the velocity of propagation of sound is 331.4 m/s. At 22° Celsius, from equation 2.6, the velocity of propagation of sound is 344.49 m/s. A sample of the transmitted and received signals in the first case is shown in appendix D. A zoomed in view of the output of the matched filter for the first case is shown in figure 5.7.

It can be seen from the tables that in each case the mean of the measured distance is within 5 centimeters of the actual distance. The standard deviation is within 14 centimeters in all three cases. It can be observed from the tables that while the individual measurements vary considerably in all the three cases taking the mean of a set of measurements averages out the variations in them. Thus in order to get an reliable measure of the range, it is very essential that the measurements be performed many times and the mean of the measurements be taken as an indication of the target distance.

Test Number	Measured Distance (meters)	Error (meters)
1	10.22	0.02
2	10.14	-0.06
3	10.19	-0.01
4	10.25	0.05
5	10.06	-0.14
6	10.18	-0.02
7	10.10	-0.10
8	10.11	-0.09
9	10.12	0.08
10	10.08	-0.12
Mean	10.15	-0.05
Standard Deviation	0.06	0.06

Table 5.1. Ranging results for a target at a distance of 10.2 meters

Test Number	Measured Range	Error
1	12.15	-0.15
2	12.22	-0.08
3	12.33	0.03
4	12.10	-0.20
5	12.36	0.06
6	12.19	-0.11
7	12.16	-0.14
8	12.27	-0.03
9	12.21	-0.09
10	12.6	0.30
Mean	12.26	-0.04
Standard Deviation	0.14	0.14

Table 5.2. Ranging results for a target at a distance of 12.3 meters.

Test Number	Measured Distance (meters)	Error (meters)
1	13.40	-0.10
2	13.51	0.01
3	13.63	0.13
4	13.35	-0.15
5	13.41	-0.09
6	13.61	0.11
7	13.54	0.04
8	13.35	-0.15
9	13.33	-0.17
10	13.39	-0.11
Mean	13.45	-0.05
Standard Deviation	0.11	0.11

Table 5.3. Ranging results for a target at a distance of 13.5 meters.

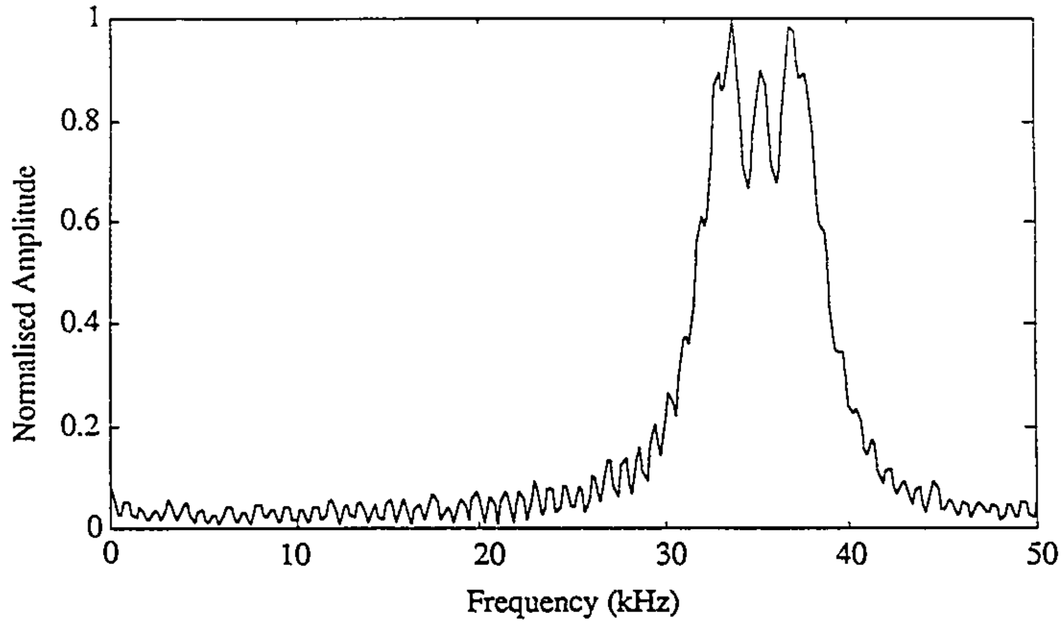

Figure 5.6. Fourier Transform of the Transmitted Chirp Pulse.

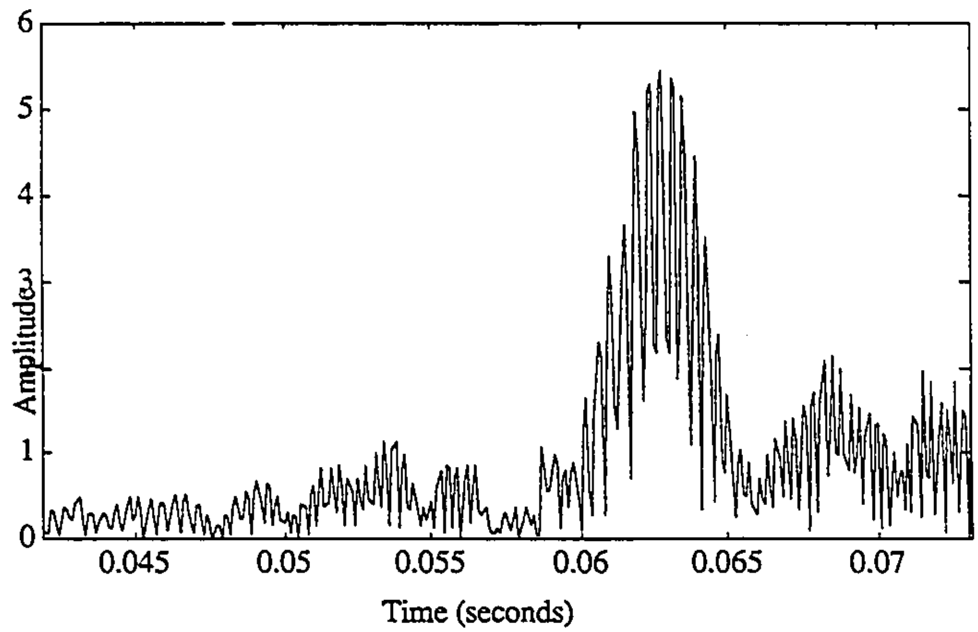

Figure 5.7. Output of the matched filter for a target a 10.2m

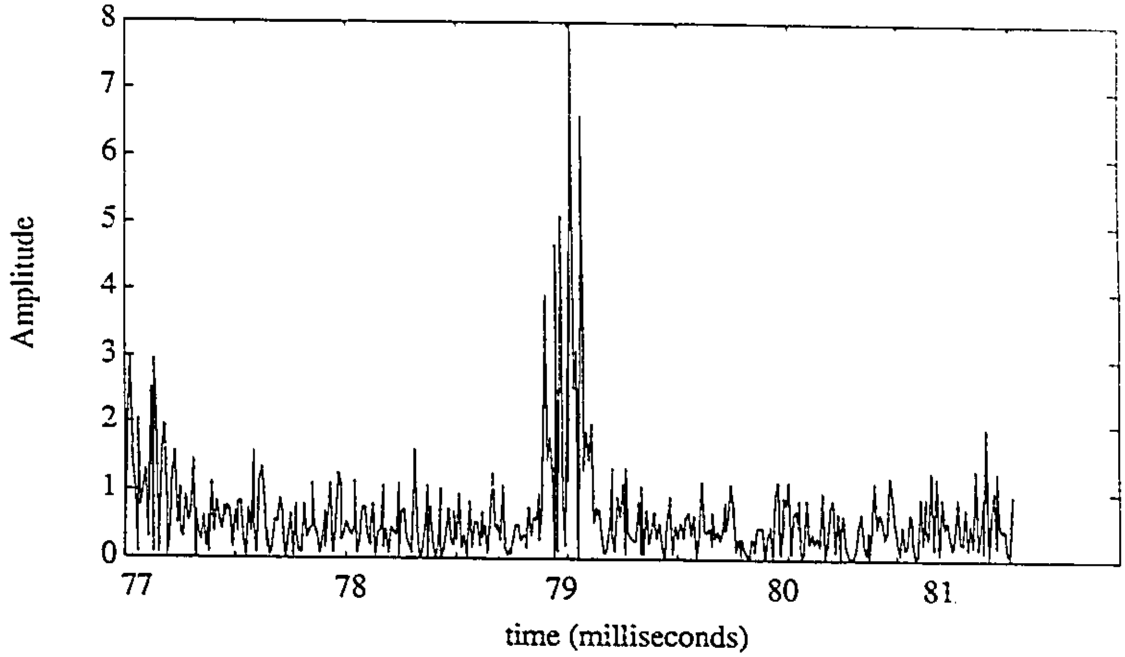

Figure 5.8 Output of matched filter for a target at 13.5 meters

VI. Conclusions and Recommendations.

6.1 Conclusions

The contributions of this study are:

1. The problem of time delay estimation as applied to ranging was simulated using both SPOX and MATLAB. Matched filters while basically used for maximizing the output signal to noise ratio, was implemented from the point of view of estimating time delay.

2. The Polaroid Ultrasonic ranging module was modified for direct interface with the Banshee board. This provides a convenient means of performing all the required operations for acoustic ranging on one host with minimum amount of interface circuitry.

3. A comparison of various window functions to reduce the temporal sidelobes present at the output of a Chirp Matched filter was performed. The reduction of sidelobe levels at the output of the matched filter is accomplished at the expense of increased main lobe width, which means that the determination of the peak would become more difficult. In terms of minimum mainlobe width increase for maximum sidelobe suppression, the Hamming window produced the best results.

4. Stream I/O routines that can be used for real-time data acquisition was implemented on SPOX. The fundamental limitation in this case was the maximum sampling frequency allowed from within the SPOX environment.

6.2 Recommendations

The present work can be extended in the following ways:

1. This study has been confined to using Linear Frequency Modulation with matched filtering. Other methods like Pseudo Noise (PN) ranging could be investigated. This would be especially interesting as the correlation functions of PN codes are two valued.

2. Repeat the present investigation with faster data acquisition rates, that is with higher sampling frequencies. In the present investigation the prime hindrance to experimenting with larger Dispersion factors was the limitation of the data acquisition system (DAS). With a faster DAS a larger proportion of the bandwidth of the transducer can be utilized.

3. Investigate using both ultrasound and microwaves together for ranging. These systems use the feature that a microwave reflection occurs only for as long as the sound wave is traveling towards the target. This feature by itself can almost double the range of an ultrasonic transducer as the sound does not have to be reflected back to the transducer in order to extract the range information. As this method does not depend on the reflected ultrasound the accuracy of this system could potentially be greater than when using ultrasound only.

4. Given the advances in VLSI technology, the matched filter can be implemented in silicon, which would provide a compact system that can be a stand alone unit without any external interface.

Appendix A

Derivation Of Matched Filter Impulse Response

This Appendix outlines the derivation of the matched filter impulse response. The idea is to choose a class of filters that would yield a maximum output in response to the input signal in the presence of additive white noise.

Let the input to the filter be $f(t)+n(t)$ where $f(t)$ is the signal and $n(t)$ is the additive white noise. The output of the filter is $f_o(t)+n_0(t)$. The ratio $\frac{|f_o(t_m)|^2}{\overline{n_0^2(t)}}$ is to be maximized.

Let the Fourier transform of $f(t)$ be $F(\omega)$ and let $H(\omega)$ be the desired transfer function of the optimum filter. Then the signal component at the output of the filter is given by :

$$f_0(t)=\frac{1}{2\pi}\int_{-\infty}^{\infty}F(\omega)H(\omega)e^{j\omega t}d\omega \tag{A.1}$$

$$f_0(t_m)=\frac{1}{2\pi}\int_{-\infty}^{\infty}F(\omega)H(\omega)e^{j\omega t_m}d\omega \tag{A.2}$$

Equation (A.2) gives the output of the filter at a time t_m , which is the best observation instant. The power spectral density of white noise is given by $S_n(\omega)=\eta/2$; thus making the output noise component

$$\overline{n_0^2(t)}=\frac{1}{2\pi}\int_{-\infty}^{\infty}\frac{\eta}{2}|H(\omega)|^2d\omega$$

(A.3)Substituting (A.3) and (A.2) into $\frac{|f_o(t_m)|^2}{\overline{n_0^2(t)}}$,

$$\frac{|f_o(t_m)|^2}{\overline{n_0^2(t)}}=\frac{\left|\int_{-\infty}^{\infty}F(\omega)H(\omega)e^{j\omega t_m}d\omega\right|^2}{\pi\eta\int_{-\infty}^{\infty}|H(\omega)|^2d\omega} \tag{A.4}$$

The Schwartz inequality is given as:

$$\left|\int_{-\infty}^{\infty} f_1(x) f_2(x) dx\right|^2 \le \int_{-\infty}^{\infty} |f_1(x)|^2 dx \int_{-\infty}^{\infty} |f_2(x)|^2 dx \qquad \text{(A.5)}$$

The above holds good if and only if $f_1(x) = kf_2^*(x)$ where k is an arbitrary constant.

Comparing equations (A.5) and the numerator of equation (A.4) ,

$$\left|\int_{-\infty}^{\infty} F(\omega) H(\omega) e^{j\omega t_m} d\omega\right|^2 \le \int_{-\infty}^{\infty} |H(\omega)|^2 d\omega \int_{-\infty}^{\infty} |F(\omega)|^2 d\omega \qquad \text{(A.6)}$$

Substituting (A.6) in (A.4)

$$\frac{|f_o(t_m)|^2}{\overline{n_0^2(t)}} \le \frac{1}{\pi\eta} \int_{-\infty}^{\infty} |F(\omega)|^2 d\omega \qquad \text{(A.7)}$$

$$\left.\frac{|f_o(t_m)|^2}{\overline{n_0^2(t)}}\right|_{\max} = \frac{E}{\eta/2} \qquad \text{(A.8)}$$

Where E is the energy in $f(t)$ for a one Ohm load. However from Schwartz in equality equation (A.8) hold s good if and only

$$H(\omega) = kF^*(\omega) e^{-j\omega t_m} \qquad \text{(A.9)}$$

Taking the inverse Fourier transform of the above yields the impulse response of the optimum fitter,

$$h(t) = kf^*(t_m - t) \qquad \text{(A.10)}$$

From equation A.10 it is obvious that the desired impulse response of the optimum filter is the mirror image of the desired signal $f(t)$, delayed by an interval t_m. Since the impulse response of such a filter is matched to the signal the term MATCHED FILTER arises. Substituting Equation (A.9) in (A.2) it is clear that at the instant of time t_m, the output of the matched filter depends only on the total energy content in the signal. Equation (A.9) states that the matched filter attenuating those strongly those frequency components that have little signal energy while attenuating very little those components that have high signal energy.

Appendix B

Program Listing for Matched Filter Simulation

In SPOX.

```
#include <spox.h>
#include <math.h>
#include <stdio.h>
#include <asig.h>
#include <aspi.h>
#include <time.h>
#ifndef M_PI
#define M_PI 3.14159265
#endif
#define LEN    4096
#define ARRAYSIZE LEN*sizeof(Float)
#define TABLESIZE ((3*LEN)/4)*sizeof(Float)


Void smain()
{
SV_Vector cvec,mag,fvec,temp,table,conjfvec,res,dvec ;
SV_Cursor vcur,vcub,vcuc ;
SV_View view;
SIG_Stream input;
SIG_Attrs sig_attrs;
FILE *ou,*im;
Float aa,xt,*bb,count;
Int len,alp,flag;
Float error,t,a;
time_t tt;
```

```
table=SV_create(FLOAT, SA_create(SG_SRAM,TABLESIZE,NULL), NULL);
fvec=SV_create(FLOAT, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
res=SV_create(FLOAT, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
cvec=SV_create(COMPLEX, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
dvec=SV_create(FLOAT, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
mag=SV_create(COMPLEX, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
temp=SV_create(FLOAT, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
SV_table(table,SV_FOURIER);
ou=fopen("out.m","w");
fprintf(ou,"%s","a=[");
 im=fopen("iout.m","w");
 fprintf(im,"%s","b=[");
 SV_fill(fvec,0.0);
 srand((unsigned)time(&tt));
 count=0.0;
                              /* Generate  The Echo*/
for(len=SV_scan(fvec, &vcur);len>0;len--)
{

    if((len>=2048+256)&&(len<2048+256+512))
    {
    t=(Float)(count)/(100000.0);
    count=count+1.0;
    *(Float *) SV_next(&vcur)=cos(2.0*M_PI*(30000.0*t+1000000.0*t*t));
    fprintf(im,"%f\n",cos(2.0*M_PI*(30000.0*t+1000000.0*t*t)));
    }
    if((len<2048+256)||(len>=2048+512+256))
    {
```

```
      *(Float *) SV_next(&vcur)=0.0;
    fprintf(im,"%f\n",0.0);
      }
}           count=0.0;
            xt=1.0;

for(len=SV_scan(dvec, &vcuc);len>0;len--)
{           flag=1;
      if((len<=4096)&&(count<512))
      {     flag=0;
      t=(Float)(count)/(100000.0);
      count=count+1.0
```

/* Modify The next line If usage of Any Windows For side lobe reduction are necessary*/

```
*(Float *) SV_next(&vcuc)=cos(2.0*M_PI*(30000.0*t+1000000.0*t*t));
 if(count>=512) flag==1;
      }
      xt=xt+1.0 ;
      if((flag==1))
      {
      *(Float *) SV_next(&vcuc)=0.0;
      }
 }
 SV_fft(fvec,cvec,table);
 SV_fft(dvec,mag,table);
 SV_conj(mag,mag);
```

```
SV_mul2(cvec,mag);
/* SV_abs(cvec,dvec); */
SV_ifft(mag,temp,table);

/* ((Complex *)SV_loc(mag, 0))->imag=0.0 ; */

for(len=SV_scan(temp, &vcur);len>0;len--)
{
fprintf(ou,"%f\n ",((*(Float *) SV_next( &vcur))));
}

fprintf(ou,"%s","];");
fprintf(im,"%s","];");
fclose(ou);
fclose(im);
SV_assign(temp,res);


SV_divs(temp,temp,255.67);
SV_abs(temp,temp);
SV_muls(temp,temp,2.0);

SV_assign(temp,res);

    /* The Following sends the output after Matched
                filtering to the D/A */
  sig_attrs=SIG_ATTRS;
   sig_attrs.sample_rate=100000;
   sig_attrs.file_data_type=SIGF_C30float;
   sig_attrs.buffer_type=SIGB_Float;
   sig_attrs.file_data_range=1.0; sig_attrs.buffer_range=1.0;
```

```
for(alp=0;alp<200;alp++)
{ SV_assign(res,temp);
input=SIG_vector_open("ad16:0",SIG_WRITE,temp,NULL);
SIG_getattrs(input,&sig_attrs); SIG_vector_put(input,temp);
SIG_close(input);
}
}
```

Appendix C

Program Listing For Data to and From Ranging Module

```
/* Due to memory management problems on the system used, the matched filtering is*/
/*done offline */
#include <spox.h>
#include <math.h>
#include <stdio.h>
#include <asig.h>
#include <aspi.h>
#include <time.h>
#ifndef M_PI
#define M_PI 3.14159265
#endif
#define LEN   7000
#define ARRAYSIZE LEN*sizeof(Float)
#define TABLESIZE ((3*LEN)/4)*sizeof(Float)


Void smain()
{
SV_Vector fvec,initala,initalb,res,temp,finalc ;
SV_Cursor vcur,vcub,vcuc ;
SV_View view;
SIG_Desc input,output;
SIG_Attrs sig_attrs;
FILE *ou,*im;
Float aa,xt,*bb,count,sq;
Int len,alp,flag;
Float error,t,a,xx;
time_t tt;
```

```
initalb=SV_create(FLOAT, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
initala=SV_create(FLOAT, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
finalc=SV_create(FLOAT, SA_create(SG_SRAM,ARRAYSIZE,NULL), NULL);
res=SV_create(FLOAT, SA_create(SG_SRAM,32*sizeof(Float),NULL), NULL);
temp=SV_create(FLOAT, SA_create(SG_SRAM,32*sizeof(Float),NULL), NULL);
fvec=SV_create(FLOAT, SA_create(SG_SRAM,4096*sizeof(Float),NULL), NULL);

im=fopen("inp.m","w");

for(len=SV_scan(temp, &vcur);len>0;len--)
{
   t=(Float)(len-1)/(100000.0);

        xx=1.0+cos(2.0*M_PI*(30000.0*t+1000000*t*t));

      if(xx>0.0) sq=1.0;
      else if(xx<0.0) sq=-1.0;

      sq=sq+1.0;
      *(Float *) SV_next(&vcur)=sq;
      fprintf(im,"%f\n",sq);
}
fclose(im);

SV_fill(initala,1.0);

     sig_attrs=SIG_ATTRS;
```

```
sig_attrs.sample_rate=100000;
sig_attrs.file_data_type=SIGF_C30float;
sig_attrs.buffer_type=SIGB_Float;

   SV_assign(initala,initalb);
      /* The Following piece of code Sends out The INIT
                    Signal*/

input=SIG_vector_open("ad16:1",SIG_WRITE,initala,NULL);
SIG_getattrs(input,&sig_attrs); SIG_vector_put(input,initala);
SIG_close(input);
SV_assign(temp,res);

                    /* End INIT */

/*      The Following Sends out The Chirp      */
output=SIG_vector_open("ad16:0",SIG_WRITE,temp,NULL;
SIG_getattrs(output,&sig_attrs); SV_assign(res,temp);
SIG_vector_put(output,temp);
SIG_close(output);

                    /*   End Chirp  */

      /* The Following Reads in the Echo   */

 input=SIG_vector_open("ad16:0",SIG_READ,fvec,NULL);
SIG_getattrs(input,&sig_attrs); SIG_vector_get(input,fvec);
SIG_close(input);

                    /* End Echo */
}
```

```
/* Write To Disk */

ou=fcpen("out.m","w");

fprintf(ou,"%s","a=[");

for(len=SV_scan(fvec, &vcur);len>0;len--)

{

fprintf(ou,"%f\n", *(Float *) SV_next(&vcur));

}

fprintf(ou,"%s","];");

fclose(ou);

}
```

Appendix D

Plots of Oscilloscope Waveforms

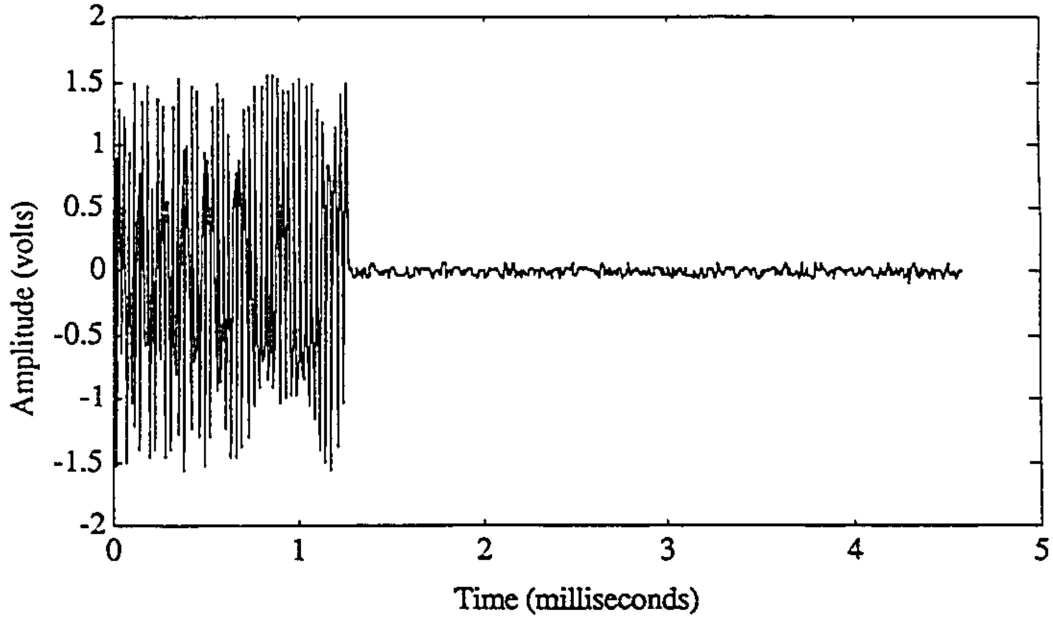

Figure D.1. Oscilloscope trace of the transmitted signal.

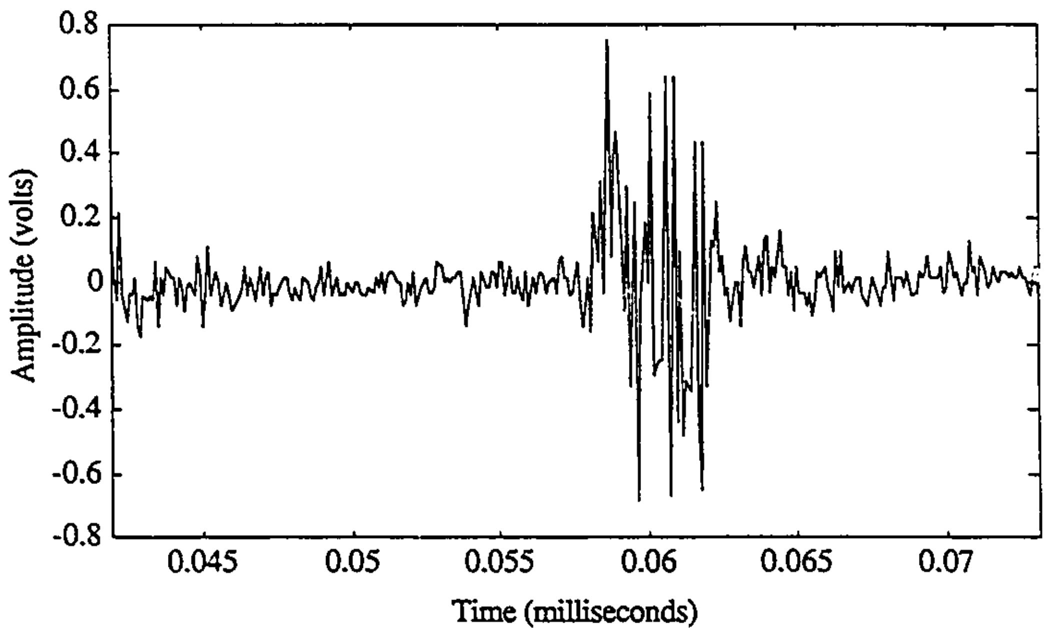

Figure D.2 Oscilloscope trace of the received echo.

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